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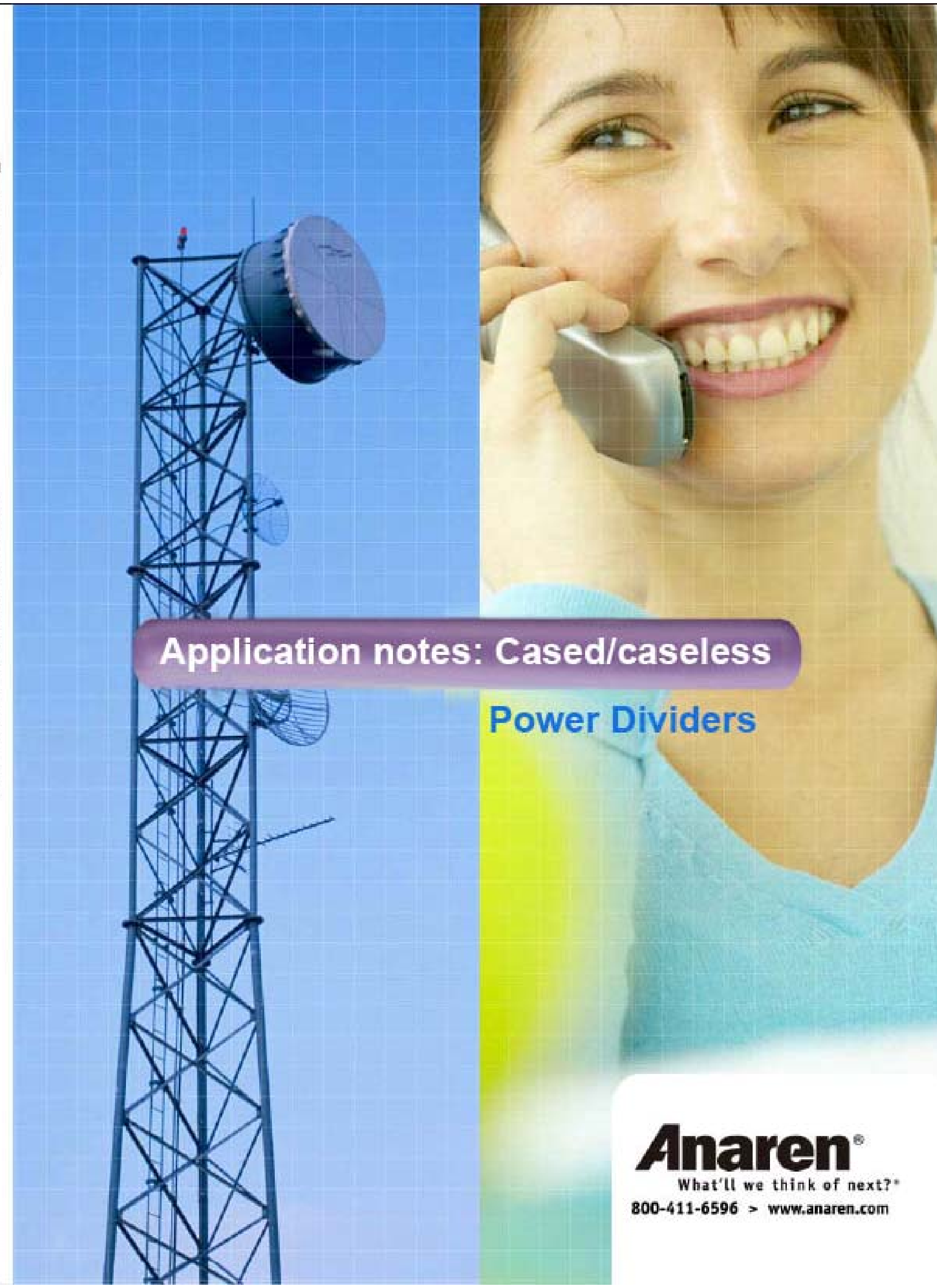

Application notes: Cased/caseless

Power Dividers

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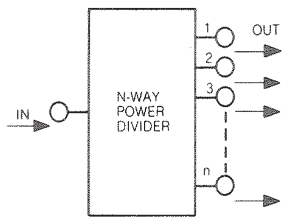
Although both in-phase (Wilkinson) and quadrature (90°) hybrid couplers may be used for coherent power divider/combiner applications, fundamental differences exist making each more suitable for specific applications. The following note discusses the technical properties of these devices.

Basic N-Way Power Dividers

Figure 1 - Basic Power Divider



An n-way power divider is shown in Fig. 1. The device has a single input port and n output ports. Ideally, input power would be divided equally between the output ports. The output phase relationship would depend upon the construction of the device. If the device were an in-phase divider, the output ports would be in phase. In a quadrature hybrid divider, the output ports would have a 90° (multiple of 90°) phase relationship.

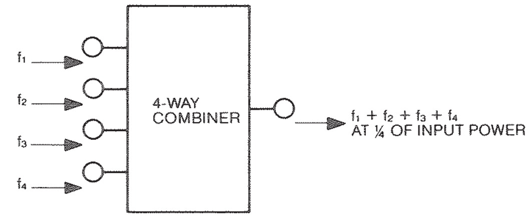
Using Dividers As Combiners

These devices, either quadrature hybrid or in-phase dividers, can be used as coherent combiners as well as dividers providing the reciprocity of the device is understood. For lossless recombination, the same amplitude and phase relationships which exist at the output of the device when used as a divider must drive the n input ports of the device when used as a combiner. A divider will losslessly combine n input signals providing they are of the proper input phase and amplitude relationship. If this is not the case, power is lost in the combiner.

In many applications coherent addition of signals is not a requirement. An example of this is where n signals of different frequencies are applied to a device with a single output port (multiplexer). One wishes to see the sum of the signals at the output. This may be accomplished using hybrid or in-phase combiners provided losses can be tolerated. Using an n-way combiner in this manner, only 1/n of the input power will appear at the output.

A four-way combiner used as a multiplexer is shown in Fig. 2. Four signals at four different frequencies of unit power level each are applied at the inputs. The output is the sum of the four signals but each has ¼ unit of power. The signals have been combined, but at the expense of a loss in power. Lossless multi-plexing can only be done with filter networks.

Figure 2 - 4-Way Combiner as a Multiplexer

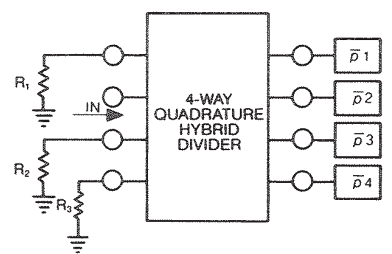


Quadrature (90°) Hybrid Power Dividers

Anaren Application Note, "90° Hybrids for Transistor Power Amplifiers," (P. 86) discusses the match and transfer properties of quadrature hybrid couplers. The conclusions drawn are that poorly matched devices may be placed at the output ports of a hybrid divider without deteriorating the input match to the divider. In a sense, then, the hybrid divider acts like an isolator provided the devices have nearly identical reflection coefficients.

Four-way Divider

Figure 3 - 4-Way Quadrature Hybrid-Divider with Reflection Coefficients, $\bar{\rho}$



The case for a four-way hybrid power divider is illustrated by Fig. 3. The divider is terminated by four devices having complex reflection coefficients, $\bar{\rho}_1$, $\bar{\rho}_2$, $\bar{\rho}_3$, and $\bar{\rho}_4$. An expression for the input reflection coefficient of the divider is given by:

$$\bar{\rho}_{in} = \tfrac{1}{4}\left(-\bar{\rho}_1 + \bar{\rho}_2 - \bar{\rho}_3 + \bar{\rho}_4\right)$$

It can be seen that if $\bar{\rho}_1 = \bar{\rho}_2 = \bar{\rho}_3 = \bar{\rho}_4$, then $\bar{\rho}_{in} = 0$. The reflected power from the mismatch at the output goes to the terminations, R_1, R_2 and R_3 as can be seen by the expression for the power to the loads:

$$P_{R1} = \frac{P_{in}}{8}\left(\bar{\rho}_1 + \bar{\rho}_2\right)^2$$

$$P_{R2} = \frac{P_{in}}{8}\left(\bar{\rho}_3 + \bar{\rho}_4\right)^2$$

$$P_{R3} = \frac{P_{in}}{8}\left(\bar{\rho}_1 - \bar{\rho}_2 + \bar{\rho}_3 - \bar{\rho}_4\right)^2$$

Note: R_3 is the "isolated" termination for the input hybrid in the 4-way quadrature hybrid divider.

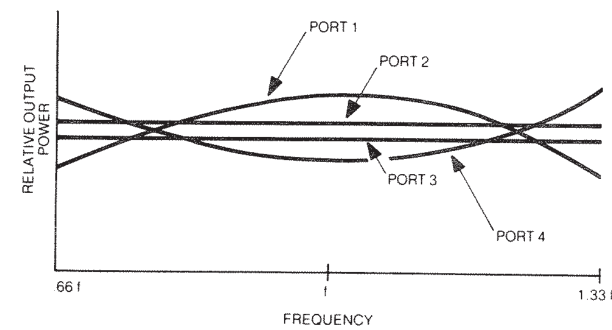
Power Handling

Another advantage of the quadrature hybrid divider is power handling capability. The terminations may be brought out of the device so that high power terminations may be used (i.e., relatively largely heat-sink or finned loads.). The stripline circuitry has a power handling capability in excess of 200 watts CW. In any given application, the power handling is usually limited by the loads. The table on page 86 gives expressions for power dissipated in terminations for hybrid dividers and combiners.

Amplitude Balance

The primary disadvantage of quadrature hybrid dividers is power imbalance between output ports. The imbalance which occurs for a four-way divider over an octave bandwidth is shown in Fig. 4. As can be seen, two ports track closely while the other two diverge at the band edges and at the center of the band (fc). Specifications for a four-way divider are usually ±1 dB for an octave bandwidth device. Better results, of course, can be achieved over narrow bandwidths with optimization of the divider.

Figure 4 - Typical Output Characteristics: Octave Band 4-Way Quadrature Divider



Phase Balance

A possible disadvantage of a quadrature hybrid divider (for certain applications) is that the output ports are not in phase. The ports have a 90° (or a multiple of 90°) phase relationship which remains constant over octave bands. Specifications for hybrid dividers are for the relative phase variation from 90° (or multiple of 90°) over the frequency range specified.

Binary Outputs

Quadrature dividers can only be constructed with a binary number (2^n) of output ports (i.e. 2, 4, 8, 16). The final section of this note discusses the construction of n-way power dividers.

In-Phase (Wilkinson) Power Divider

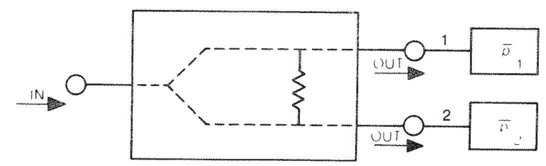
The in-phase (Wilkinson) power divider has the advantages of excellent output port amplitude balance (over octave or wider bandwidths) and in-phase power division. The power split and phase balance is theoretically perfect and nearly ideal results can be achieved in practice.

Power Handling

The chief disadvantage (for power applications) is that the terminations cannot be brought out externally to the device. A constraint on the internal terminations is that the terminations must be much less than a wavelength in any dimension. This constraint limits worst case power handling of some devices to approximately 200 milliwatts. These are "fail safe" specifications. The final power handling capability depends upon the external terminations of the divider. As an example, a two-way divider with terminations is illustrated having complex reflection coefficients $\bar{\rho}_1$ and $\bar{\rho}_2$. The power (P) lost in the internal termination is given by:

$$P = \frac{P_{in}}{4}\left(\bar{\rho}_1 - \bar{\rho}_2\right)^2$$

Figure 5 - 2-Way In-Phase Divider



It can be seen that if the reflection coefficients are nearly identical no power will be lost in the internal termination. The worst case is for one reflection coefficient to be 180° out of phase with the other. In this case, all reflected power is dissipated in the internal termination. A good example is if both terminals are open ($\bar{\rho}_1 = \bar{\rho}_2 = 1.0$) or short ($\bar{\rho} = \bar{\rho}_2 = -1.0$). In this case no power is dissipated in the external termination. If, however, one output port is open circuited and the other short circuited, ($\bar{\rho}_1 = 1.0$, $\bar{\rho}_2 = -1.0$) then all of the input power is dissipated in the internal termination.

Figure 5a - Input Power (P_{in}) for 2-Way In-Phase Power Dividers.

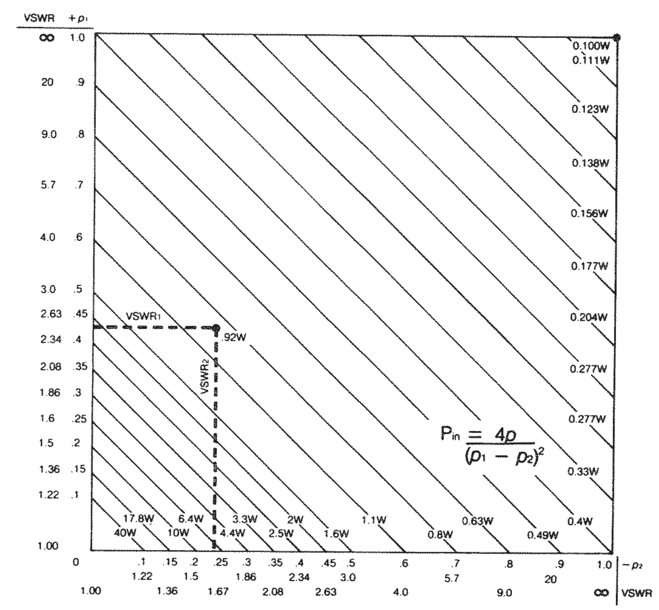


Figure 5A is useful for determining the power handling capabilities of the 2-way, in-phase power dividers of Figure 5.

Worst case phase (180°) is assumed for the external terminations. Internal load dissipation of only 100 mW is assumed (to provide a safety factor).

Example:

VSWR of 2.5:1 is expected at port 1 and 1.6:1 at port 2. The graph shows that the 2-way, in-phase power divider can accept approximately 0.92 W at the input port under these conditions.

Input Match

Unlike the quadrature hybrid power divider, if the output ports are loaded with devices of nearly identical reflection coefficients, the input match degrades according to the magnitude of the reflection coefficients. This is seen by an expression for the input reflection coefficient.

$$\overline{\rho}_{in} = \tfrac{1}{2}(\overline{\rho}_1 + \overline{\rho}_2)$$

Unlike the quadrature hybrid divider, if the outputs are open or short circuited, ($\overline{\rho}_1 = \overline{\rho}_2 = 1.0$ or $\overline{\rho}_1 = \overline{\rho}_2 = -1$) the input reflection coefficient is unity.

N-Way Outputs

In-phase power dividers, like the quadrature hybrid dividers are most easily constructed with a binary (2^n) number of output ports. N-way dividers can be constructed but are most easily done in cylindrical geometries (more difficult from a construction point of view) than the binary dividers which lend themselves to planar geometries.

In-Phase Power Dividers as Power Combiners

The in-phase power divider may be used for combining RF signals. There are two types of signal combining:

a) Coherent signal combining, where the signals have exactly the same period or frequency, but not necessarily the same amplitude or phase code.

b) Non-coherent signal combining, where the combined signals do not have the same period, and may not have the same amplitude.

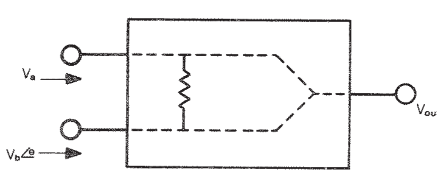
Coherent Signal Combining

The output of a 2-way in-phase power divider (see Figure 6) when used as a coherent signal combiner can be determined by the following equation:

$$V_{out} = \tfrac{1}{\sqrt{2}}\sqrt{(V_a + V_b \cos\theta)^2 + (V_b \sin\theta)^2}$$

where V_a and V_b are coherent input voltages and θ is the phase difference between V_a and V_b.

Figure 6 - 2-Way Power Divider Used As Coherent Signal Combiner



Non-coherent Signal Combining

Combining two non-coherent signals will cause 3 dB loss for each input signal in addition to the losses normally incurred when the device is used as a power divider. This is because the signals are asynchronous and will not correlate. Therefore, the power divider output will be $1/\sqrt{2}\ V_{in}$ for each input signal.

Quadrature and In-Phase Combinations

Both quadrature hybrid and in-phase power dividers have common properties of good match at all ports and high isolation between output ports. Devices may be constructed using combinations of both types of power dividers to gain the advantages of both. For example, an eight-way divider can be constructed from in-phase and quadrature hybrid dividers. Using this approach, input match under matched load conditions can be guaranteed and the divider would have better amplitude balance than a device made up of quadrature hybrid dividers alone.

In power applications (i.e. amplifiers), in-phase dividers can be kept in the low power region of the device with hybrids in the higher power region. Again, better amplitude balance is obtained while maintaining the advantage of good input match.

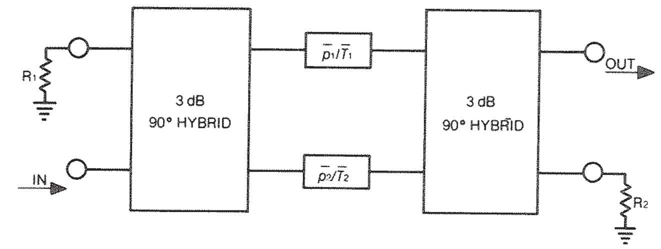
Finally, where output phase is not important, n-way dividers, for odd division or even division other than 2^n, may be done using any combination of quadrature hybrid dividers, in-phase power dividers, and backward wave couplers. In this manner, 3, 5, 6, . . . way dividers have been constructed for particular applications.

Input Match and Power Relationships

The following tables have expressions for input reflection coefficients, output power and power lost to terminations for two-way and four-way quadrature hybrid and in-phase combiner/dividers. The expressions are in terms of reflection and transmission coefficients for devices placed between the divider and combiner. From these tables, one can also obtain data for dividers and combiners and combiners taken by themselves.

For example, as dividers, one only need be concerned with the reflection coefficients. As combiners, one can start with the amplitude and phase at the input ports which will give perfect combination. Then by progressive perturbation of phase and amplitude at any one of the ports, one can determine power output and power lost to any one of the loads.

Two-Way Quadrature (90°) Hybrid Combiner—Dividers



Input Reflection Coefficient $= \tfrac{1}{2}(\overline{\rho}_1 - \overline{\rho}_2)$

Power Output $= \frac{P_{in}}{4}(\overline{T}_1 + \overline{T}_2)^2$

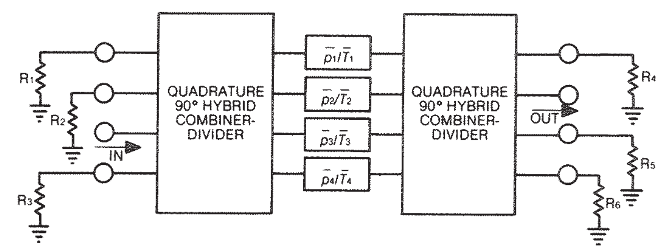
Power Lost in $R_1 = \frac{P_{in}}{4}(\overline{\rho}_2 + \overline{\rho}_1)^2$

Power Lost in $R_2 = \frac{P_{in}}{4}(\overline{T}_2 - \overline{T}_1)^2$

$\overline{\rho}$ = Voltage Reflection Coefficients

$\overline{T}$ = Voltage Transmission Coefficients

Four-Way Quadrature (90°) Hybrid Combiner-Dividers



Input Reflection Coefficient $= \tfrac{1}{4}(\overline{\rho}_1 - \overline{\rho}_2 + \overline{\rho}_3 - \overline{\rho}_4)$

Power Output $= \frac{P_{in}}{16}(\overline{T}_1 + \overline{T}_2 + \overline{T}_3 + \overline{T}_4)^2$

Power Lost in $R_1 = \frac{P_{in}}{8}(\overline{\rho}_1 + \overline{\rho}_2)^2$

Power Lost in $R_2 = \frac{P_{in}}{16}(\overline{\rho}_1 - \overline{\rho}_2 + \overline{\rho}_3 - \overline{\rho}_4)^2$

Power Lost in $R_3 = \frac{P_{in}}{8}(\overline{\rho}_3 + \overline{\rho}_4)^2$

Power Lost in $R_4 = \frac{P_{in}}{8}(\overline{T}_1 - \overline{T}_2)^2$

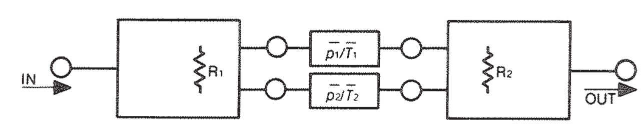
Power Lost in $R_5 = \frac{P_{in}}{16}(\overline{T}_1 + \overline{T}_2 - \overline{T}_3 - \overline{T}_4)^2$

Power Lost in $R_6 = \frac{P_{in}}{8}(\overline{T}_3 - \overline{T}_4)^2$

$\overline{\rho}$ = Voltage Reflection Coefficients

$\overline{T}$ = Voltage Transmission Coefficients

Two-Way In-Phase Combiner-Dividers



Input Reflection Coefficient $= \tfrac{1}{2}(\overline{\rho}_1 + \overline{\rho}_2)$

Power Output $= \frac{P_{in}}{4}(\overline{T}_1 + \overline{T}_2)^2$

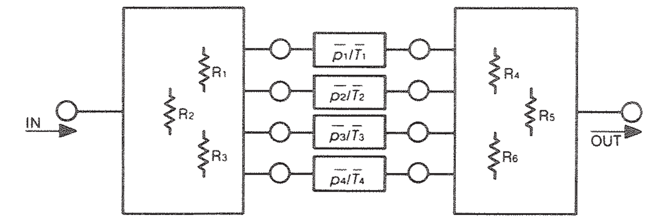
Power Lost in $R_1 = \frac{P_{in}}{4}(\overline{\rho}_1 - \overline{\rho}_2)^2$

Power Lost in $R_2 = \frac{P_{in}}{4}(\overline{T}_1 - \overline{T}_2)^2$

$\overline{\rho}$ = Voltage Reflection Coefficients

$\overline{T}$ = Voltage Transmission Coefficients

Four-Way In-Phase Combiner-Dividers



Input Reflection Coefficient $= \tfrac{1}{4}(\overline{\rho}_1 + \overline{\rho}_2 + \overline{\rho}_3 + \overline{\rho}_4)$

Power Output $= \frac{P_{in}}{16}(\overline{T}_1 + \overline{T}_2 + \overline{T}_3 + \overline{T}_4)^2$

Power Lost in $R_1 = \frac{P_{in}}{8}(\overline{\rho}_1 - \overline{\rho}_2)^2$

Power Lost in $R_2 = \frac{P_{in}}{16}(\overline{\rho}_1 + \overline{\rho}_2 - \overline{\rho}_3 - \overline{\rho}_4)^2$

Power Lost in $R_3 = \frac{P_{in}}{8}(\overline{\rho}_3 - \overline{\rho}_4)^2$

Power Lost in $R_4 = \frac{P_{in}}{8}(\overline{T}_3 - \overline{T}_2)^2$

Power Lost in $R_5 = \frac{P_{in}}{16}(\overline{T}_1 + \overline{T}_2 - \overline{T}_3 - \overline{T}_4)^2$

Power Lost in $R_6 = \frac{P_{in}}{8}(\overline{T}_3 - \overline{T}_4)^2$

$\overline{\rho}$ = Voltage Reflection Coefficients

$\overline{T}$ = Voltage Transmission Coefficients

90° Hybrids for Transistor Power Amplifiers

Recent advances in RF power transistors have made it technically feasible and economic to replace tube-type amplifiers with solid state power amplifiers. At this time, power amplifiers operating at 1kW (CW) over the UHF band 225-400 MHz are entirely feasible. The more common power requirements in this frequency range, however, are in the 10 W to 100 W range.

Design Problems

Design problems for wideband amplifiers center around two main areas:

1) The higher the power rating of an RF transistor, the lower and more complex is the input impedance. This makes it difficult to match the device successfully over any bandwidth.

2) The cost and reproductibility of the transistor increases non-linearly with the power rating. Increased power handling capability is usually achieved by dc paralleling many low power stages on a single chip. Maintaining an even power division over an increasingly larger number of stages is both a difficult materials problem and a fabrication feat. This increases the probability that a few stages will be stressed beyond capability, and the transistor is destroyed.

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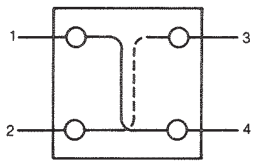
Using the 3 dB, 90° Hybrid

Techniques for overcoming these difficulties have been described by Benjamin [1,2] and others. These techniques are based primarily on the use of quadrature (90°) hybrids (3 dB couplers).

The 90° hybrid (Fig. 1) is a reciprocal four-port device that behaves like a pair of matched tees with common output ports, and isolated input ports (hybrid tee). The two tees are different only in the relative phasing of the voltages in the output ports. A signal into port 1 divides equally between ports 3 and 4 and ideally, port 2 is perfectly isolated. The voltage at port 4 lags the voltage at port 3 by 90° (hence, the term quadrature).

Figure 1 - 3 dB, 90° Hybrid



A signal into port 2 also divides equally between ports 3 and 4 with port 1 isolated, but in this case the voltage at port 4 leads the voltage at port 3 by 90°. Since the device is reciprocal, the same situation holds true when ports 3 and 4 are used as the isolated input ports.

If ports 3 and 4 are terminated in mismatches with voltage reflection coefficients $\overline{\rho}_3 - \overline{\rho}_4$) respectively and a signal is incident on port 1, then the voltage reflection cofficient at port 1 is given by:

$$\tfrac{1}{2}\,(\overline{\rho}_3 - \overline{\rho}_4)$$

and at port 2

$$\tfrac{1}{2}\,(\overline{\rho}_3 + \overline{\rho}_4)$$

Thus, if port 2 is terminated in a matched load, the hybrid behaves like a tee with an input reflection coefficient half the difference between the reflection coefficients of the output loads.

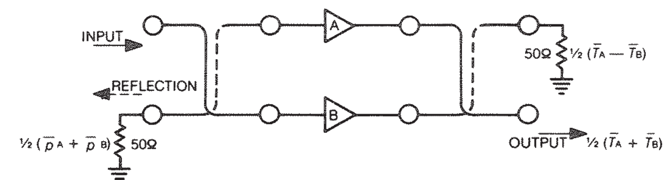
If the output reflection coefficients are equal ($\overline{\rho}_3 = \overline{\rho}_4$), then the input port is perfectly matched. Note that this is true even for such drastic output conditions as open or short circuits.

Thus, if the output ports are terminated in equal mismatches the 90° hybrid behaves like a tee where the reflected power never reaches the input port.

When any two circuit components (transistors, for example) having complex voltage transmission coefficients $\overline{T}_A$ and $\overline{T}_B$ (and reflection coefficients $\overline{\rho}_A$, $\overline{\rho}_B$) are placed between two hybrids, the voltage transmission of the complete network is: $\tfrac{1}{2}\,(\overline{T}_A + \overline{T}_B)$

(1) J. Benjamin, "RF Power Combinations Using Hybrid Junctions," and "Broadband Transistor Power Amplifier Concept," ITT Semiconductors.
(2) J. Benjamin, "Use Hybrid Junctions for More VHF Power," Electronic Design, August 1, 1968.

Figure 2 - Balanced Amplifier



The voltage is $\tfrac{1}{2}\,(\overline{T}_A - \overline{T}_B)$ and is absorbed in the termination of the isolated output port.

When the transmission coefficients are identical ($\overline{T}_A = \overline{T}_B$) then none of the transmitted signal is lost to the isolated port and all appears at the output. When the input reflection coefficients ($\overline{\rho}_A$, $\overline{\rho}_B$) are identical the input hybrid sums all the reflected voltage $\tfrac{1}{2}\,(\overline{\rho}_A + \overline{\rho}_B)$ to the termination at the isolated input port.

The new network of Figure 2 is a balanced amplifier having low input VSWR and output power twice that of a single transistor.

Wideband Power Amplifiers

Now let us consider some of the problems of building a wideband power amplifier. The RF transistors have low complex input impedances and some attempt at matching must be made to realize the gain potential. To obtain flat gain response across the band, however, it is not necessary to match the transistor completely across the band. Since the maximum available gain decreases with frequency (approximately 6 dB per octave), the matching network can be chosen to maximize the gain at the high frequency end of the band and allow the mismatch to increase towards the lower end of the band. The reflection coefficient of such an amplifier can be as high as 0.8 at the lower end of an octave band. However, combining two such amplifiers by means of a pair of quadrature hybrids allows the realization of a matched power amplifier module with a flat gain response.

Using this module as a building block, high power octave bandwidth amplifiers can be constructed (paralleling modules) by means of the same quadrature hybrid technique.

In some instances, four or even eight modules could be combined in the last power stage. To facilitate construction and lower the cost, Anaren can provide four-way combiner divider networks for selected bands.

Many power transistor manufacturers have developed application information on this subject [3,4]. The effects of hybrid characteristics (phase and amplitude balance) on balanced amplifier performance are discussed at length on page 88.

(3) J. Johnson, et al., "Solid Circuits," (Chapter 16), Application Book 2.2.8.0A Communications Transistor Company, San Carlos, Calif.
(4) Application Notes AN-6010, AN-6118, AN-6126; RCA Solid State Division, Sommerville, N.J.

Effect of Hybrid Phase and Amplitude Balance on Balanced Amplifier Performance

90° Hybrids For Balanced Amplifiers

The Ultra-Miniature series of 3 dB, 90° hybrids is specifically intended for installation directly in the microstrip circuitry employed for most balanced transistor amplifer designs. Thirty-four standard models in 15 case styles cover the frequency ranges from 30 MHz to 6.0 GHz. The popular 225-400 MHz band is covered by 4 standard case styles.

A balanced transistor stage usually consists of two electrically similar transistors whose inputs and outputs are combined in 3 dB, 90° hybrid couplers. The characteristics of the hybrids provide an amplifier module with good impedance match at input and output while simultaneously giving good phase and amplitude characteristics. (Anaren Application Note, "90° Hybrids for Transistor Power Amplifiers," discusses the match and transfer properties of quadrature hybrid couplers more completely.)

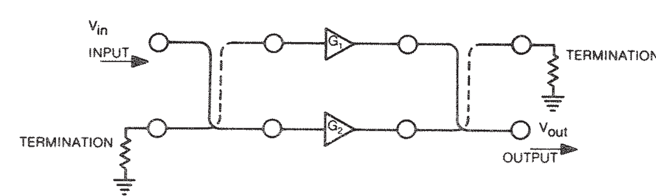
Using a balanced amplifier module as a building block, high power amplifiers can be constructed by paralleling modules using the same quadrature hybrid technique. Anaren offers 4 models of 4-way Combiner-Dividers designed specifically for this application; these devices use the same lightweight construction and microstrip solder terminals as the Ultra-Miniature 90° hybrids.

Hybrid Phase and Amplitude Balance

The balanced amplifier module provides good input and output match. It also provides good power handling ability (up to 1,000 W CW for specially constructed units) because the terminated ports can be brought out of the unit and high power terminations used. A disadvantage of the module is the power imbalance between coupled ports of the quadrature hybrids, which results in some power lost at the output port. This is not as serious a problem as it might first appear.

The effects of the amplitude and phase balance of the hybrids on balanced amplifier performance can be evaluated using Figure 1.

Figure 1 - Balanced Amplifier



Let the input be $V_1 = 1$ volt and G_1, G_2 be the voltage gains for the two transistors.

The output voltage can be shown to be:

$$\frac{V\ out}{V\ in} = V\ out = e^{-j(\phi - \frac{\pi}{2})}\,[(G_1 + G_2)\sin\theta\cos\theta] \quad (1)$$

Where: ϕ = total insertion phase from input to output
θ = coupling angle
$V_{in} = 1$

θ is a function of the length of the coupled section of the hybrid, the propagation constant and the coupler geometry. A narrow band hybrid that has exactly –3 dB coupling to each output at midband has $\theta = 45°$.

Amplitude Performance of "Perfect" Hybrids

Evaluating equation (1) at midband for $\theta = 45°$:

$$V\ out = e^{-j(\phi - \frac{\pi}{2})}\,[(G_1 + G_2)(.707)(.707)]$$

$$= e^{-j(\phi - \frac{\pi}{2})}\,[(G_1 + G_2)(.5)] \quad (2)$$

Assume $G_1 = G_2 = 1$ for ease of computation: equation (2) becomes

$$V\ out = e^{-j(\phi - \frac{\pi}{2})}\,[(2)(.5)]$$

$$= e^{-j(\phi - \frac{\pi}{2})}\,[1.00]$$

Letting $\phi = 0$ (no phase imbalance):

$$V\ out = e^{-j\frac{\pi}{2}} \quad (1)$$

In other words, an amplifier module using "perfect" hybrids (no phase or amplitude imbalance) delivers all its power to the output port and the output has a phase of –90° relative to the input.

Amplitude Performance of Octave-Band Hybrids

A normal octave band hybrid is specified with a maximum differential power split at its outputs of 1 dB (amplitude balance of $\pm$0.5 dB) and could therefore have a coupling angle $\theta = 48.5°$:

Evaluating equation (1) at midband for $\theta = 48.5°$.

$$V\ out = e^{-j(\phi - \frac{\pi}{2})}\,[(G_1 + G_2)(.749)(.663)]$$

$$= e^{-j(\phi - \frac{\pi}{2})}\,[(G_1 + G_2)\ .496] \quad (3)$$

Assuming $G_1 = G_2 = 1$ equation (3) becomes

$$V\ out = e^{-j(\phi - \frac{\pi}{2})}\,[(2)\ .496]$$

Letting $\phi = 0$ (no phase imbalance)

$$V\ out = e^{-j\frac{\pi}{2}}\,(.992)$$

V out = .992 of the input voltage (at a relative phase of –90°)

V out = –.07 dB

Therefore, a coupler imbalance of ± 0.5 dB results in a power loss of only .07 dB at the amplifier output.

Note: Even if the amplitude balance was ± 1.0 dB, the power loss would only be:

$$V\,out = e^{-j(\phi - \frac{\pi}{2})}\,[(G_1 + G_2)(\sin 51.5^\circ)(\cos 51.5^\circ)]$$

$$= 2\,(.782)(.623)$$
$$= .974$$
$$= -0.2\text{ dB}$$

Phase Performance of Hybrids

The power lost due to a phase imbalance of 3° ($\pm$1.5°) for each hybrid in the amplifier can most easily be calculated by assuming that the imbalance is due to an additional line length of 6° (let the couplers and amplifiers be "perfect"). This assumes the phase errors are adding up in a worst case fashion:

Then $\phi = 180^\circ + 6^\circ = \pi + 6^\circ$

$$V\,out = e^{-j(\pi + 6^\circ - \frac{\pi}{2})}\,[(1 + 1)(\sin 45^\circ)(\cos 45^\circ)]$$

$$= e^{-j(\frac{\pi}{2} + 6^\circ)}[(2)(.707)(.707)]$$

$$= e^{-j(96^\circ)}\,[1]$$

$$= (\cos 96^\circ - j \sin 96^\circ)[1]$$

$$= (-\sin 6^\circ - j \cos 6^\circ)[1]$$

$$V\,out = (-.1045 - j\,.995)$$

The output voltage vector at a relative phase of –90° is now equal to .995 of the input.

V out = –.04 dB

Hybrid phase imbalance of $\pm$1.5° (adding in worst case fashion) results in a power loss of only .04 dB at the amplifier output. Please note that the phase imbalance of most Anaren hybrids is *typically* $\pm$0.5° and not the maximum spec of $\pm$1.5°.

Summary

Balanced transistor amplifiers using 3 dB hybrids provide good input and output impedance match and offer good power handling capability.

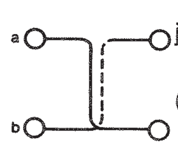
An amplifier using "perfect" hybrids (no amplitude or phase imbalance) will focus all the power to the output port. Any amplitude or phase imbalance will cause some power to be directed to the terminated port and will result in power lost at the output port.

The effects of amplitude and phase imbalance for an amplifier using octave band hybrids were examined. Only 0.11 dB is lost to defocusing: about 0.07 dB for amplitude imbalance and 0.04 dB for phase imbalance.

A closing comment: Better hybrid amplitude balance could, in fact, be provided by using multi-section hybrids. However, production tolerances would make it impossible to economically provide octave-band amplitude balance better than $\pm$0.2 dB. More importantly, any amplifier performance improvement due to better amplitude balance is usually nullified by the increased insertion loss of the multi-section hybrid.

Derivation of Output Expressions for 90° Hybrids

The general expression for any backward wave 90° hybrid coupler with two inputs (a, b) is:



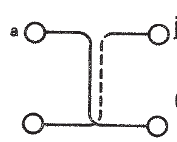
$$j\,(a \sin\theta)\,e^{-j(\beta\ell + \epsilon)} + (b\cos\theta)\,e^{-j(\beta\ell + e)}$$

$$(a\cos\theta)\,e^{-J(\beta\ell + \epsilon)} + j(b \sin\theta)\,e^{-j(\beta\ell + \epsilon)}$$

where: $(\beta\ell + \epsilon)$ is the coupler insertion phase plus a small dispersive phase error

and: β = propogation constant $(\frac{2\pi}{\lambda})$
ℓ = coupled length
ϵ = small phase dispersion error term
θ = coupling angle and a, b are voltages

The output expressions simplify if a is the only input (b = 0):



$$j\,(a \sin\theta)\,e^{-j(\beta\ell + \epsilon)}$$

$$(a \cos\theta)\,e^{-j(\beta\ell + \epsilon)}$$

Letting a = 1 (volt) and ignoring ϵ:

$$V_{coupled} = j\,(\sin\theta)\,e^{-j\beta\ell}$$

$$V_{dc} = (\cos\theta)\,e^{-j\beta\ell}$$

These expressions can be simplified even further at midband where:

$$\beta\ell = \frac{\pi}{2}\;(\ell = \frac{\lambda}{4})$$

then $e^{-j\beta\ell} = \cos\frac{\pi}{2} - j\sin\frac{\pi}{2} = 0 - j = -j$

and $V_{coupled} = j\,(\sin\theta)(-j) = \sin\theta$

$V_{dc} = (\cos\theta)(-j) = -j\cos\theta$

If $\theta = 45^\circ$ (perfect 3 dB coupling)

$V_{coupled} = \sin 45^\circ = 1/\sqrt{2}$

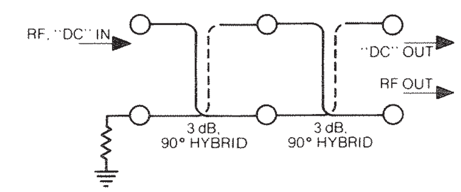
$V_{dc} = -j\cos 45^\circ = -j/\sqrt{2}$

Signal Diplexer ("Crossover") Using 3 dB 90° Hybrids

When two 3 dB, 90° hybrids are connected in tandem, the the resulting four-port network (Fig. 1) displays some interesting properties. This occurs because two couplers connected back-to-back behave like a single coupler with a coupling angle equal to the sum of the two individual coupling angles. If each of the two couplers have maximum coupling angles of 45° (3 dB couplers) their combined coupling angle is 90° (a 0 dB coupler). This device is useful as a diplexer and can be used to separate low frequency signals (IF, video, dc) from high frequency (RF) signals.

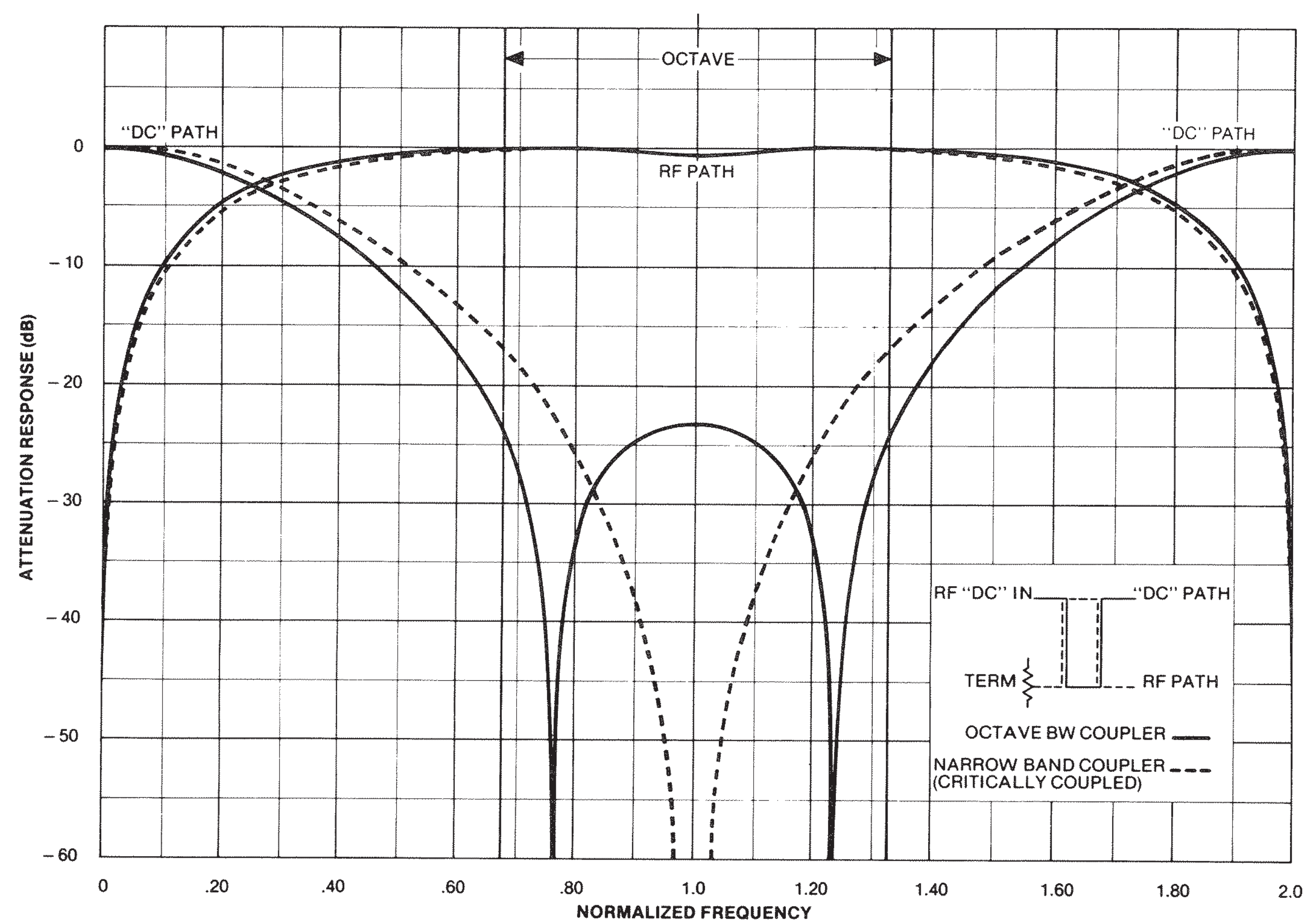
Figure 2 is a graph showing the attenuation response of the network to signals of different frequencies. (1) The frequency scale is normalized to the hybrid's band center (fc = 1.0) to make the graph usable for all tandem connected 3 dB,

Figure 1 - Signal Diplexer ("Crossover")



90° hybrids (having the same model number). The attenuation scale is dB loss from the RF, "DC" input (ignoring coupler insertion loss).

Figure 2 - Diplexer ("Crossover") Response

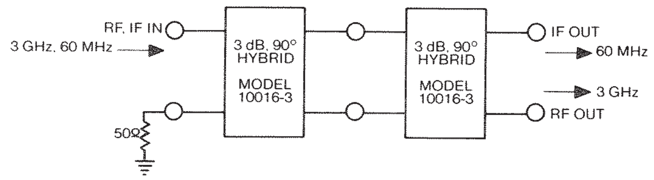


(1) The solid curves on Figure 2 show the response when using typical octave band devices. The dashed curve shows the response when using critically-coupled hybrids. i.e. devices that have 3 dB coupling only at band center. The difference in attenuation response characteristics between the octave and narrow-band devices is most evident when looking at the low frequency ("DC") path. The nulls in the attenuation response occur at points of exactly 3 dB coupling: band center (f_C) for the critically coupled hybrids and at the crossovers (.77f_C & 1.23f_C) for the octave band hybrid. Production variations that affect the crossover (3 dB) points will affect the position and depth of the nulls.

Using the Graph

Assume that a transmission line contains signals at 3 GHz and 60 MHz. Two 3 dB, 90° hybrids (Anaren Model 10016-3) can be used to direct the low frequency (60 MHz) signal to one output port and the 3 GHz signal to the other output, as shown by Fig. 3.

Figure 3 - RF/IF Diplexer



The normalized freq. of 1.0 (fc) on the chart corresponds to the hybrid *center* frequency (3.0 GHz for the Model 10016-3). The chart shows that there is negligible loss between the RF, IF IN port and the RF OUT port for the 3 GHz signal; the 60 MHz signal (a normalized frequency of .02) is down approximately 24 dB at the RF OUT port.

The 60 MHz signal exits at the "DC" port with negligible loss while the 3 GHz signal is down approximately 24 dB. (If the coupler was critically coupled, meaning it was designed for exactly 3 dB coupling at band center, the rejection at the "DC" port to the 3 GHz signal would be infinite.)

The diplexer in the example just cited could be used for increasing the LO/IF or RF/IF isolation of an S-band mixer having a 60 MHz IF. If the mixer alone had 20 dB LO/IF isolation, the diplexer would provide at least 24 dB *additional* rejection at any point in the 2-4 GHz octave. In addition, the LO and RF signals would be terminated in a matched load and would not reflect back into the IF port to cause additional intermodulation problems. In this case, the diplexer performs as a low-loss, low-cost filter, matched at all frequencies.

Extended Band Coupling Data for Single-Section Hybrids and Directional Couplers

The coupled power from both the dc (mainline) and ac (coupled) ports of single-section 3 dB and directional couplers is frequency sensitive. The charts and examples in this section are designed to provide out-of-band coupling data for these devices.

Figure 1 shows typical extended-band characteristics for an octave-band 3 dB, 90° hybrid. The frequency scale is normalized to 1.0 at the coupler center frequency (f_c).

Example 1: A model 10016-3 hybrid is to be used at 1200 MHz rather than its octave band design range of 2-4 GHz. What are the coupling values at the 0° and –90° ports?

First; calculate f_c for the 10016-3:

$$f_c = \frac{f_{LO} + f_{HI}}{2} \quad \frac{2\text{ GHz} + 4\text{ GHz}}{2} = 3\text{ GHz}$$

Second; normalize 1200 MHz to f_c: $\frac{1200}{3000} = 0.4$

Third; read the coupling values from the two curves at $f_c = 0.4$:

Coupling value at coupled (0°) arm = –5.45 dB
Coupling value at dc (–90°) arm = –1.45 dB

Figure 1 - Extended Band Coupling Data For Octave Band 3 dB, 90° Hybrids

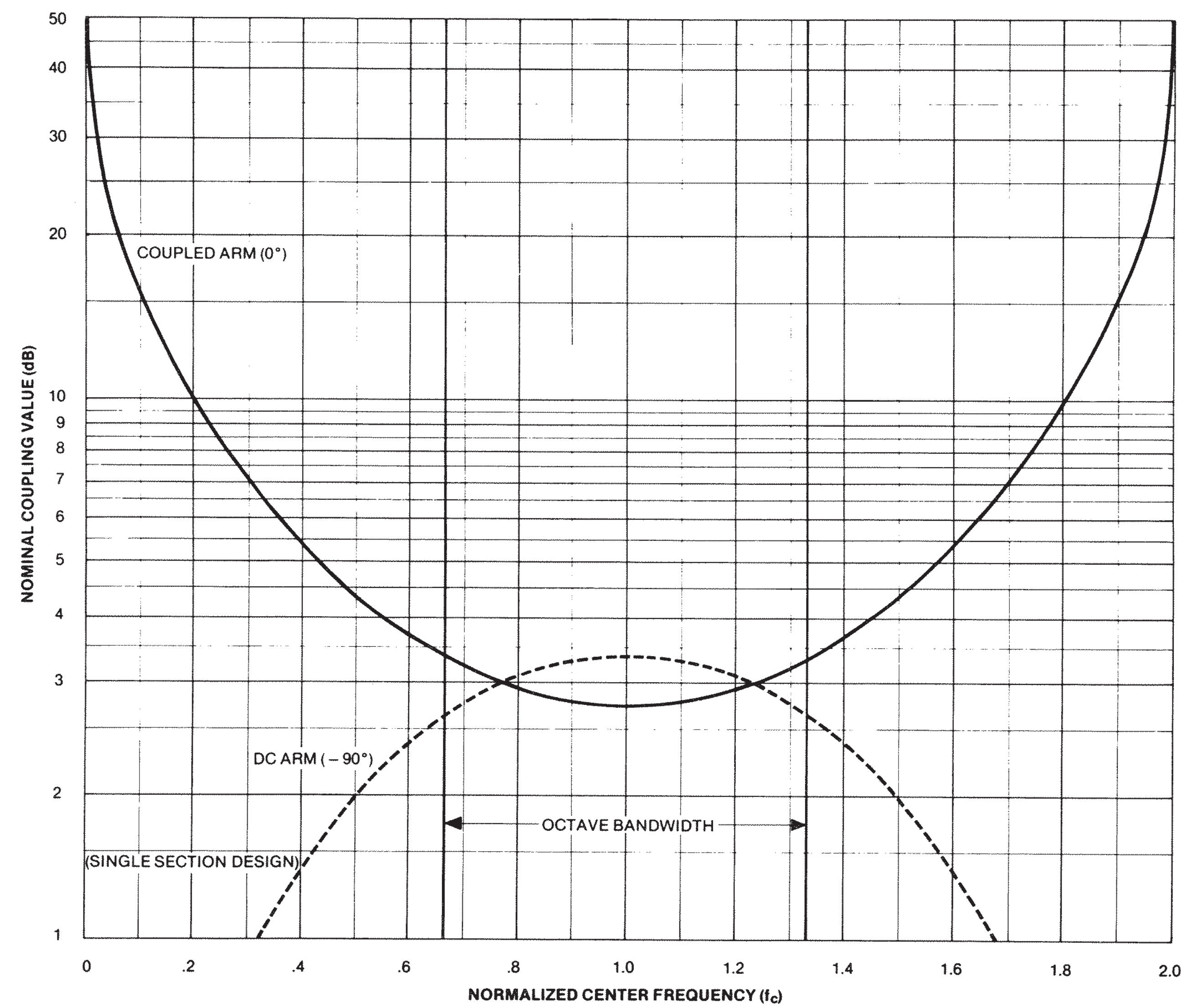


Figure 2 is similar to Figure 1 except that it details the frequency response of the *coupled arm only* for couplers of various coupling values.

Example 2: What is the expected roll-off at 5 GHz for a Model 10616-6 directional coupler?

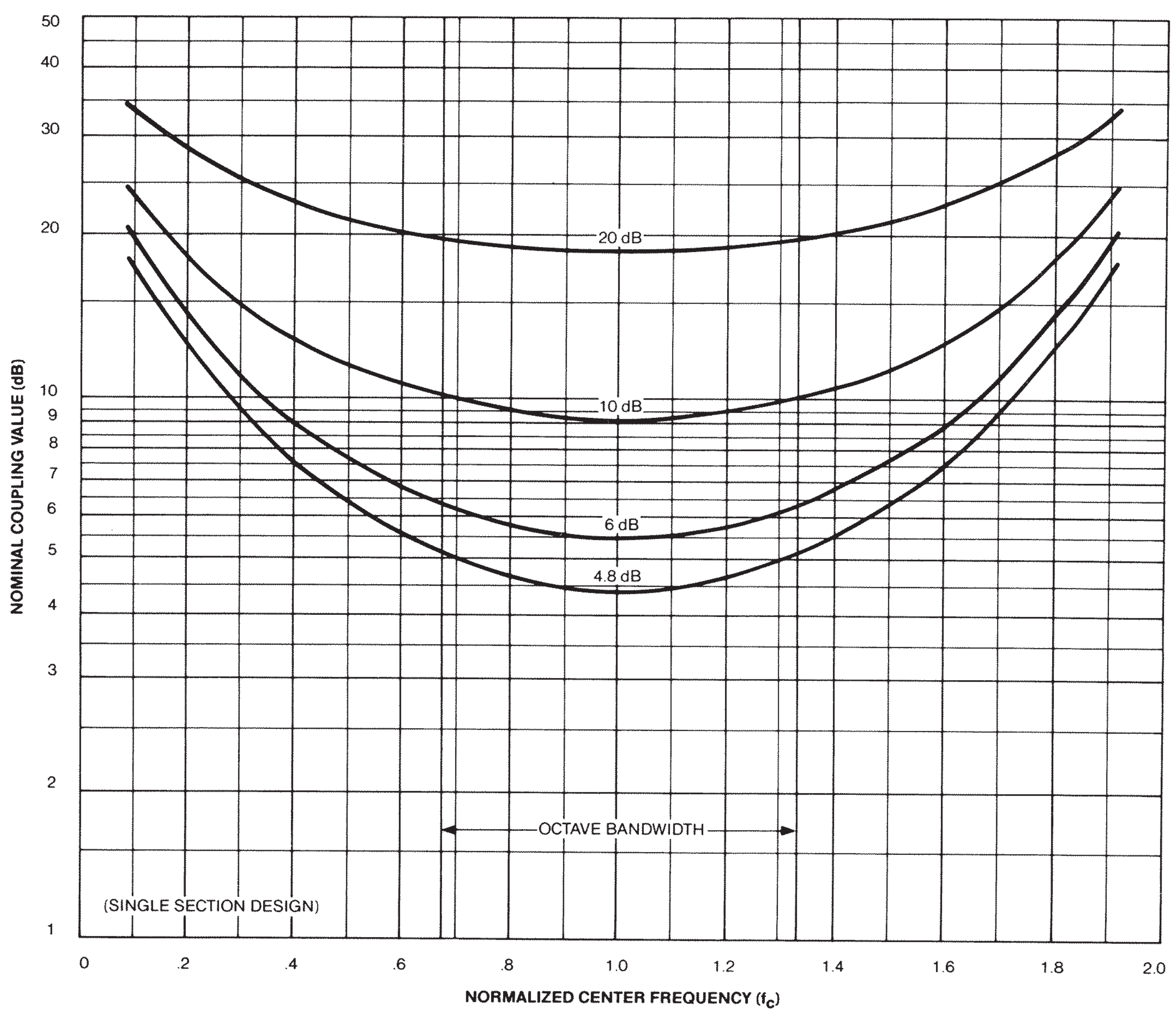
First: calculate f_C for the 10616-6:

$$f_C = \frac{f_{LO} + f_{HI}}{2} = \frac{2\text{ GHz} + 4\text{ GHz}}{2} = 3\text{ GHz}$$

Second; normalize 5 GHz to $f_C \frac{5}{3} = 1.666$

Third; read the coupling value from the –6 dB curve at $f_C + 1.666$: approx. –10.2 dB

Figure 2 - Extended Band Coupling Data for Directional Couplers



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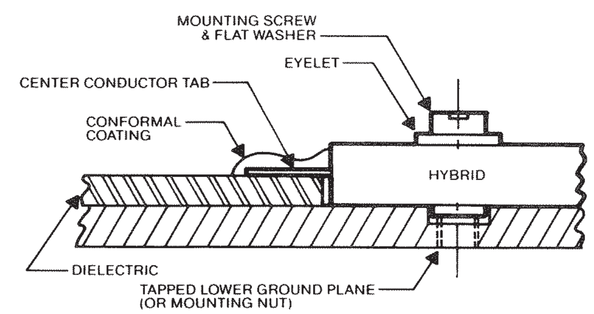
Installation Details for Caseless Couplers and Combiner-Dividers

Anaren's caseless couplers *must* be installed with the label up for good case to ground plane contact. Caseless couplers can be installed in microstrip or stripline transmission media. Most units are designed for circuits using .030 inch dielectric material. Note that the ground plane mounting surface must be counterbored to clear the eyelet protrusion on the caseless coupler as shown in Figure 1.

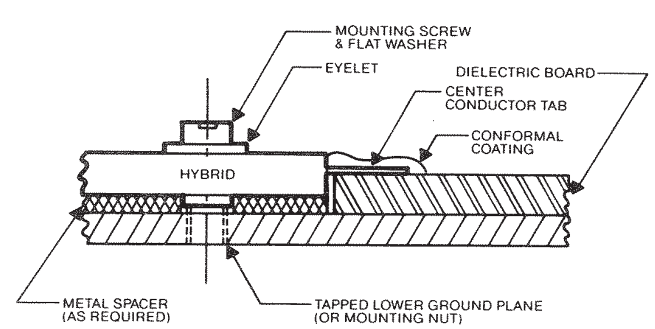
Figure 1 – Basic Mounting, Side View



This counterbore assures that the coupler is well grounded to provide the extremely low inductance ground paths necessary for good high frequency performance. It also promotes heat dissipation in high power applications and ensures flush contact of the coupler conductor tabs to the mating microstrip conductors.

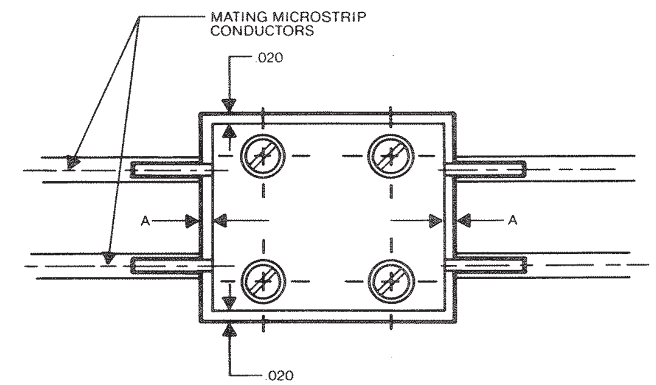
Figure 2 shows the use of a metal spacer to provide good electrical ground contact the and flush tab contact when the dielectric board is more than one-half the hybrid thickness.

Figure 2 – Mounting with Spacer, Side View



In normal installation, an area is cut-out of the dielectric board to accommodate the coupler as shown in figure 3.

Figure 3 – Basic Mounting, Top View



To permit proper tab alignment to the board's microstrip conductors, approximately .020 inch clearance is allowed on each side of the coupler.

The clearance dimension "A" on the tab sides of the coupler should be minimized to limit the inductance caused by conductor tab spanning an air gap. An "A" dimension of .020 inch is acceptable up to 150 MHz, .010 inch up to 1 GHz and .005 inch up to 2 GHz. Above 2 GHz this dimension becomes very critical and every effort must be made to minimize it.

Screws are normally used to fasten the coupler to the system ground plane. This ground plane may be tapped or it may be drilled with clearance holes for a mounting nut.

Conductor tabs may be attached to the microstrip conductor by soldering, conductive epoxy, welding or any other low contact resistance attaching method.

High Power Considerations

For use in application at or near their rated power, Conformal coating on the tab to coupler interface is required to eliminate arcing and voltage breakdown caused by this sharp transition region. Conformal coating the tabs also is required in high humidity and high altitude application.

Heatsinking, other than normal mounting is not normally required. However, when operating at high power, any effort to improve heat dissipation will minimize the insertion loss due to copper resistance increasing as temperature increases.

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