Wide-Band Lumped-Element Quadrature 3-dB Couplers in Microstrip

D. P. Andrews and C. S. Aitchison, *Fellow, IEEE*

***Abstract*—A design procedure for wide-band quadrature 3-dB couplers at microwave frequencies is presented in this paper, wherein lumped elements are incorporated into a microstrip circuit. Design examples for an octave-band coupler and a one–and-one-half octave-band coupler are presented. The topologies presented offer a convenient and cost-effective solution to tight coupling problems, currently solved using distributed circuits.**

***Index Terms*—Filters, hybrid junctions, microstrip circuits, microstrip directional couplers, microwave power amplifiers, passive circuits, quadrature hybrids.**

I. Introduction

A NUMBER of solutions to the 3-dB quadrature coupler design problem at microwave frequencies are available to the engineer. These almost invariably require the construction of a tightly coupled transmission line. Design tables for the idealized circuit have been available for some time [1]. For designs up to an octave, a single-section coupler is usually sufficient. For designs much greater than an octave, a single-section coupler may still be adequate, but often it is necessary to increase the complexity to a three-section coupler, whereby the bandwidth is extended to about 4.5 : 1, for amplitude balance comparable to a single-section coupler operating over one octave. For this three-section coupler, the center section is required to be tightly coupled, with the outer sections relatively loosely coupled. When designing in microstrip, it would be most convenient if the coupling could be implemented by laying two lines side by side. However, the range of coupling values using this technique does not approach the kind of values required for 3-dB couplers, while maintaining practical circuit dimensions. One technique for overcoming this limitation is by the use of interdigitated lines, as in the Lange coupler [2]. Even this solution requires fine lines and spacing and, thus, is more appropriate where thin-film processing can be used. Lange couplers can be implemented using standard etching techniques, but the fine structure will add significantly to the printed circuit board (PCB) manufacturing cost. This is, in general, inconvenient, as it is unlikely that any other features on the PCB will be so demanding. Even after the circuit has been processed, it is still necessary to apply the bonding straps. Automatic bonding machines will ease this task, but themselves represent an expensive overhead. A particular advantage with the Lange coupler is that it can be implemented up to the highest microwave frequencies.

Owing to the difficulties experienced in solving the coupler design problem with a purely microstrip solution, it is often found convenient to use another transmission-line topology. At least one manufacturer supplies a ready made coaxial coupled line, which can be cut to the required length for a particular frequency of operation. A small selection of coupling values is available so that designs up to at least an octave can be implemented. The performance of these products is excellent. However, they are not without their shortcomings. One difficulty arises from the abrupt transition between the microstrip and coaxial lines. Little difficulty is experienced at the lower microwave frequencies, and problems only arise at frequencies in excess of about 3 GHz. Another difficulty lies in the fact that only certain coupling values are available, thus, optimizing for a particular bandwidth is not possible. Another solution to the problem of implementing a quadrature coupler is to introduce a proprietary stripline hybrid into the microstrip circuit. By this means, the designer is relieved of the problem of designing the coupler itself, and simply makes use of what is commercially available. It is relatively easy to implement tight coupling in stripline, by making use of broadside coupling. These stripline hybrids may be surface-mount or drop-in. For the surface-mount version, it is necessary for the user to implement vias in the microstrip substrate, to provide the ground connection to the hybrid. This is no problem with large-scale surface-mount technology, but is inconvenient when prototyping and in small scale production requirements. For either of these two stripline options, the designer is limited in frequency and bandwidth by what is commercially available. Custom designs may present a prohibitive cost penalty.

Quadrature 3-dB couplers that employ lumped elements in their construction have been known about for some time. The basic lumped-element coupler makes use of a perfectly coupled inductor pair, with capacitance placed between the two coupled inductors. On its own, the bandwidth of this circuit is narrow, and is only suitable for applications up to around 15%–20%. However, these narrow-band circuits may be incorporated into more complex circuits to achieve much broader bands. Cappucci invented a circuit [3] that used a pair of short transmission lines between a pair of these lumped-element narrow-band quadrature hybrids to achieve an octave band. (He also proposed a technique for implementing much larger bandwidths than this [4], but these circuits will not be considered in this paper.) This design was investigated by others [5], [6]. The same principle

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D. P. Andrews is with Vectawave Technology Ltd., Ryde, IOW P033 3QF, U.K.

C. S. Aitchison is with the Department of Electronic Engineering, University of Surrey, Surrey GU2 5XH, U.K.

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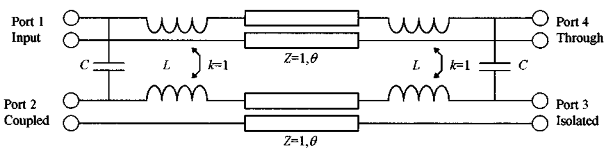


Fig. 1. Circuit diagram of the octave-band quadrature coupler.

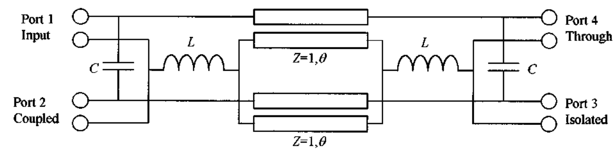


Fig. 2. Microwave version of the octave-band quadrature coupler.

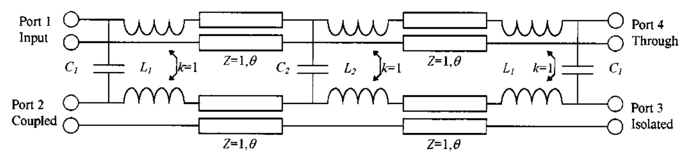


Fig. 3. Circuit diagram of the extended octave-band quadrature coupler.

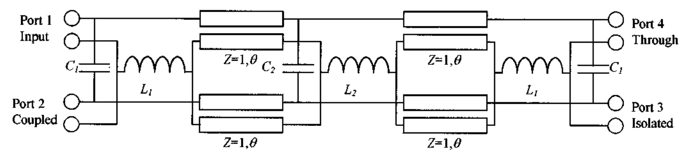


Fig. 4. Microwave version of the extended octave-band quadrature coupler.

was extended by Chen for the design of greater than an octave couplers, by the use of three lumped-element hybrids with transmission lines in between [7]. All these circuits share the same limitation in that they require the use of perfectly coupled inductors in the realization of the lumped-element hybrids. This constraint has limited the application of these circuits to rarely more than UHF frequencies. However, a new development by the authors of the lumped-element narrow-band quadrature hybrid, by which the coupled inductor is replaced by a ground line inductor, has made it possible to implement lumped-element quadrature hybrids at microwave frequencies [8]. If these microwave lumped-element quadrature hybrids are substituted in the existing designs, then it is possible to realize microwave versions of the wide-band quadrature couplers [9].

II. Prototype Electrical Design

A. Octave-Band Circuit

The prototype octave-band design as available in the literature is shown in Fig. 1. For perfect match and isolation at all frequencies, it is required that $\sqrt{L/C} = Z_0$, and that the characteristic impedances of the connecting lines is the same as the termination impedance. In addition, these constraints also ensure the through and coupled outputs to be in exact phase quadrature at all frequencies. Selection of the values of the inductors, capacitors, and connecting line electrical length can be made to optimize the circuit for a particular bandwidth. If the design is initially normalized to 1 Ω, then the inductor and capacitor values become numerically the same. For octave-band operation, centered on an arithmetic mean radian frequency of unity, data from the literature [4], [5] or direct synthesis will yield element values of

$$L \text{ and } C = 0.809$$
$$\theta = 19.5^\circ.$$

With these element values, the theoretical maximum coupling is 2.7 dB; a similar value to that obtained in the design of an optimum octave-band single-section transmission-line 3 –dB coupler.

The two lumped-element sections in Fig. 1 can be adapted, according to the technique given in [8], to remove the perfectly coupled inductors, and replace them with ground line inductors. The circuit then becomes that shown in Fig. 2. The value of the inductor is the same as the mutual inductance of the coupled inductor of Fig. 1. The values of the remaining elements also do not change.

B. Extended Octave-Band Circuit

For bandwidths much above an octave, the circuit of Fig. 1 exhibits an undesirable deterioration in amplitude balance, and the more sophisticated prototype circuit of Fig. 3, due to Chen [7], becomes a more acceptable solution. This circuit achieves acceptable performance for bandwidths up to about two octaves, and so can be considered as a suitable design for bandwidths extending beyond one octave. This circuit also exhibits ideal match, isolation and quadrature properties, given the same constraints as the octave-band circuit. For the specific case of a 3-to-1 bandwidth ratio (sometimes referred to as one-and-a-half octaves), values normalized for unity characteristic impedance and unity center radian frequency for the circuit of Fig. 3 become

$$L_1 \text{ and } C_1 = 0.278$$
$$L_2 \text{ and } C_2 = 1.704$$
$$\theta = 35.4^\circ.$$

The maximum coupling with these element values over the specified bandwidth is 2.7 dB, i.e., the same as the octave-band coupler. The lumped-element sections of the circuit of Fig. 3 should be adapted to replace the coupled inductors with ground line inductors, to make the circuit suitable for microwave operation. The circuit becomes as shown in Fig. 4.

III. Physical Implementations

A. Octave-Band Coupler

To illustrate the practicality of the octave-band circuit of Fig. 2, it was decided to design a coupler operating over a 1–2-GHz band. This modest frequency range was chosen for the purpose of proving the concept, without introducing the greater manufacturing difficulties that might be expected given a higher frequency specification.

The element values determined above for the prototype octave-band circuit of Fig. 2 can be scaled for the center frequency of 1.5 GHz and terminating impedance of 50 Ω. The element values then become

$$L = 4.29 \text{ nH}$$
$$C = 1.72 \text{ pF}.$$

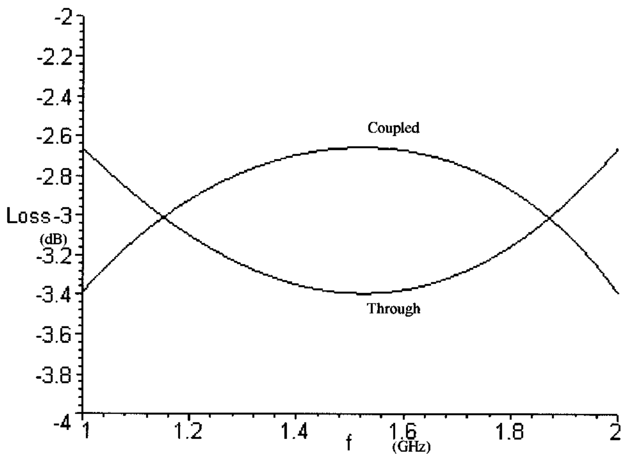


Fig. 5. Theoretical through and coupled response of 1–2-GHz coupler.

These values correspond with a simple coupler of center frequency 1.85 GHz. The interconnecting lines simply adopt the new characteristic impedance, with an electrical length of 19.5° at 1.5 GHz. With these element values substituted into the circuit of Fig. 2, the through and coupled responses become as in Fig. 5. Unlike the single-section transmission-line coupler, the theoretical response is not symmetrical with frequency. Exactly equal division occurs at 1.15 and 1.87 GHz, with the maximum coupling occurring at 1.525 GHz, just above the center frequency.

Of the circuit elements of Fig. 2, the capacitors are the most easily implemented, being simply lumped components. The ground line inductors can be implemented using the techniques of [8], whereby a pattern is etched onto the ground plane, consisting of a strip with a break in the middle, over which the microstrip lines pass. Over most of the length of the microstrip line, the ground plane strips act like normal microstrip ground. However, at the break, the ground current must find a different route. This different route is provided by allowing a current to pass along the underside of the strip to the edge of a cavity underneath the substrate, continuing its path along the surface of the cavity, and completing its path along the underside of the strip opposite the break. The current can then continue in partnership with the microstrip line, and resume a normal microstrip mode. The strip suspended over the cavity can be approximated as a wide microstrip line with air dielectric, with the two end walls of the cavity realizing short circuits. The cavity circuit thus appears, from the point-of-view of the break in the strip, as two short-circuit stubs in series. Provided the electrical length of the stubs is small at the frequency of operation, the circuit will approximate an inductor. The modeling of the cavity circuit is not accurate, owing to the discontinuities of the cavity walls, but an initial estimate can be made, which can be refined empirically. The design of the cavity circuit needs to bear in mind the requirement to fit two identical lumped-element sections in the complete coupler, with microstrip lines of sufficient length between the strip break points. In order for the cavity circuit to appear as a lumped inductor, it is desirable for it to be physically small. One way of achieving this end is to arrange a narrow ground-plane strip. This, in turn, requires the microstrip lines to be narrow (e.g., by using a thin substrate) and, hence, potentially introduce more loss. A compromise clearly has to be struck. The use of a substrate with low dielectric constant will also be preferable, as the physical length of the connecting lines will be greater, giving more flexibility in positioning the cavities. It would be desirable too for the cavity depth to be similar to its length, so as to maximize inductance for a given length/depth maximum dimension. The optimum depth may give rise to difficulties if it is too great, as it will require a thicker carrier into which the cavity must be machined.

With the previous considerations, it was determined to use a 0.79-mm-thick Arlon 25 N substrate with a relative dielectric constant of 3.28. This is greater than polytetrafluoroethylene (PTFE)-based substrates, and will yield longer connecting lines. However, this material was chosen because it is more rigid and, thus, will not deform so easily over the unsupported regions of the cavity. On this substrate, the 50-Ω connecting lines have a width of 1.8 mm, and their length needs to be 6.7 mm. A cavity depth of 5 mm was chosen, and it was determined that its length should be 11.75 mm to realize the required reactance at 1.5 GHz. The width of the cavity was set at 15 mm, though this dimension is not critical. Even with a narrow rib between the two cavities, the connecting line length still makes it necessary to move each break in the ground line away from the center of the cavity, toward the rib. As the total electrical length of the cavity circuit is only 21° at the center frequency, the inductance is not influenced significantly by the break position.

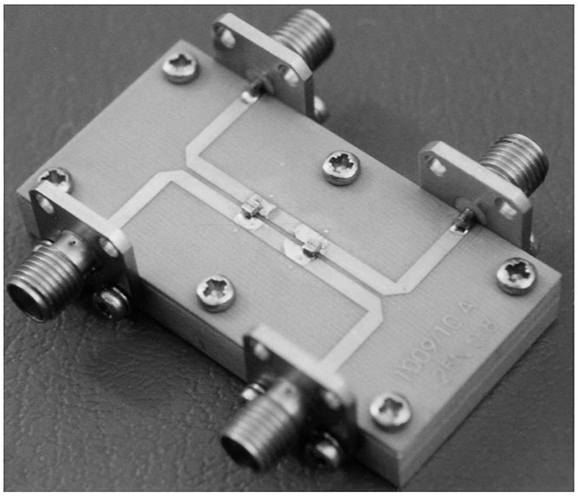
A photograph of the coupler is shown in Fig. 6. Fig. 6(a) is a top view of the complete coupler, showing the microstrip lines and lumped capacitors. SMA connectors are attached for the purpose of measurement. Fig. 6(b) is a view of the underside of the substrate, showing the ground plane etched away over the cavities, except for a 5-mm-wide strip, which itself has a break. Fig. 6(c) is a view of the aluminum carrier, showing its machined cavities.

B. Extended Octave-Band Coupler

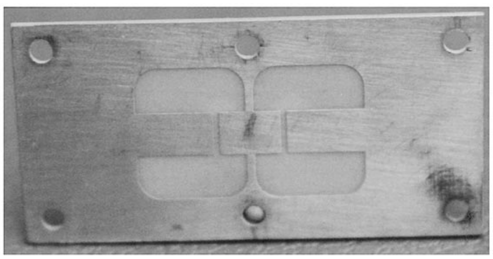
To illustrate the practicality of the extended octave-band circuit, a target frequency was taken to include the bandwidth of the octave-band circuit, and extend it to include up to 3 GHz. As in the octave-band circuit, this frequency range was chosen for the purpose of proving the concept. The element values determined above for the prototype circuit of Fig. 4 can be scaled for a center frequency of 2 GHz and terminating impedance of 50 Ω. With this transformation performed, the required element values become

$$\begin{aligned} L_1 &= 1.106 \text{ nH} \\ C_1 &= 0.442 \text{ pF} \\ L_2 &= 6.78 \text{ nH} \\ C_2 &= 2.71 \text{ pF.} \end{aligned}$$

The outside two lumped-element sections correspond to simple couplers with a center frequency of 7.2 GHz, with the center section corresponding to 1.17 GHz. Again, the interconnecting lines adopt the new characteristic impedance, with an electrical length of 35.4° at 2 GHz. The predicted through and coupled responses are shown in Fig. 7. Exact equal division occurs at three frequency points: 1.155, 2.1, and 2.875 GHz. Maximum



(a)



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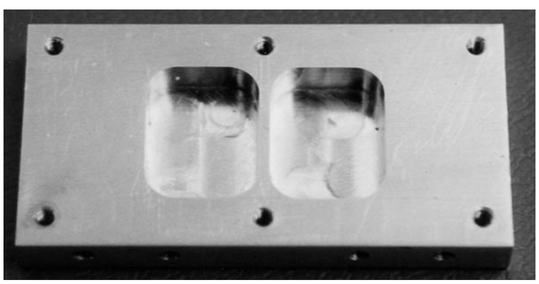


(c)

Fig. 6. (a) Complete 1–2-GHz coupler. (b) Underside of 1–2-GHZ coupler PCB. (c) 1–2-GHz coupler carrier.

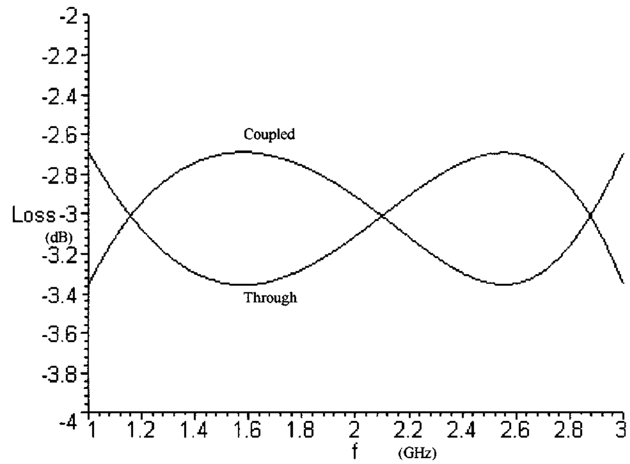


Fig. 7. Theoretical through and coupled response of 1–3-GHz coupler.

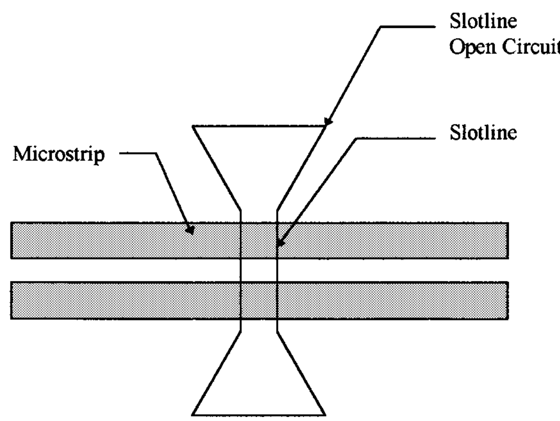
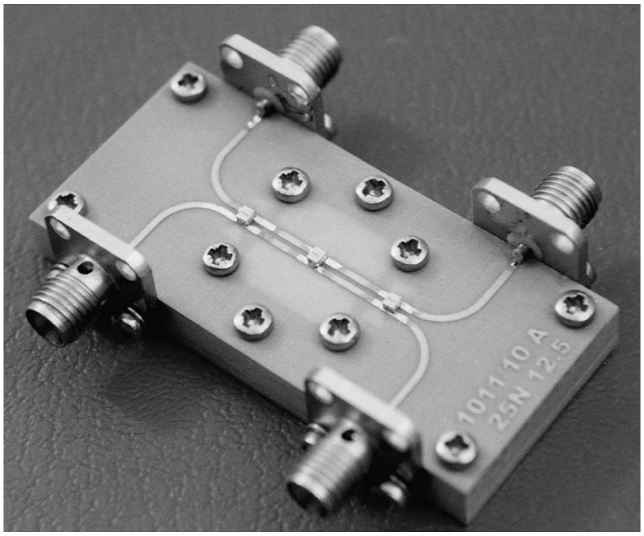


Fig. 8. "Bow-tie" ground line end section inductor in 1–3-GHz coupler.

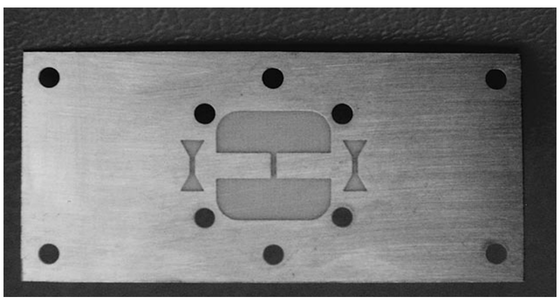
coupling occurs at 1.58 GHz, as well as the upper band edge. Minimum coupling occurs at the lower band edge and a frequency of 2.55 GHz.

Design of the circuit follows a similar procedure as for the octave-band coupler. The design of the center section inductor is more demanding, as the value is larger and has to operate to a higher frequency. As has been discussed above, the cavity circuit resembles two short-circuit stubs in series. In order to preserve a good lumped approximation over the operating frequency range, it was decided to limit the electrical length of each stub to $\lambda/16$ at 3 GHz. This equates to a physical length of 6.25 mm, the length of the cavity being twice this at 12.5 mm. The characteristic impedance of the airline stubs was then determined such that they produced the same total reactance at the center frequency of 2 GHz, as would a pure inductor of 6.78 nH. The reactance required is 42.6 Ω for each stub, and will be achieved with a characteristic impedance of 159 Ω. Microstrip synthesis gives a linewidth-to-substrate height ratio of 0.57 for this impedance. If the 5-mm strip width of the octave-band coupler was retained, then the cavity depth would need to be 8.77 mm. It was decided that this depth was too great, and that it should be limited to 5 mm, which, in turn, implies a strip width of 2.85 mm. To achieve this end, it would be necessary to choose a thinner substrate, so that 50-Ω lines could be accommodated over this strip width. A substrate thickness of 0.318 mm was available, using the same material as for the octave-band coupler, giving 0.735-mm-wide 50-Ω lines. The connecting line length to give 35.4° at 2 GHz was determined to be 9 mm.

The end-section inductors present a different problem, in that their value is small. To implement them, the technique discussed in the design of the 6.4-GHz coupler in [8] was used, whereby they are implemented as a pair of slotline open circuits, with a short length of slotline between them The open circuits are operated well below their normal frequency range, to exploit their low-frequency inductive properties. The structure resembles a "bow-tie," and is shown in Fig. 8. The slotline length and width was chosen to be the same as in [7], which is 2-mm long and 0.5-mm wide. The open-circuit length needed to be a little less than in that reference, to shift the operating frequency from 6.4 GHz to the present requirement of 7.2 GHz. To achieve this, the side angle was retained at 30°, and the length was reduced from 2 to 1.77 mm. In order to prevent a short circuit of the



(a)



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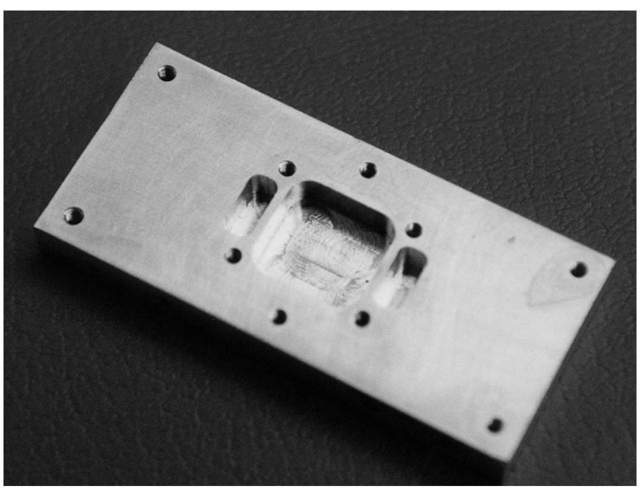

(c)

Fig. 9. (a) Complete 1–3-GHz coupler. (b) 1–3-GHz coupler PCB. (c) 1–3-GHz coupler carrier.

"bow-tie" inductor, a cavity is necessary underneath. This was made large enough so as not to influence the inductance. The complete coupler is shown in Fig. 9. Fig. 9(a) shows the top circuit, with three lumped-element capacitors defining the length of the interconnecting lines. Fig. 9(b) shows the underside of the substrate, showing the different means of implementing the inductors of the circuit. Fig. 9(c) is the aluminum carrier, showing the three cavities. It can be seen that the design only just permits sufficient room between the three cavities.

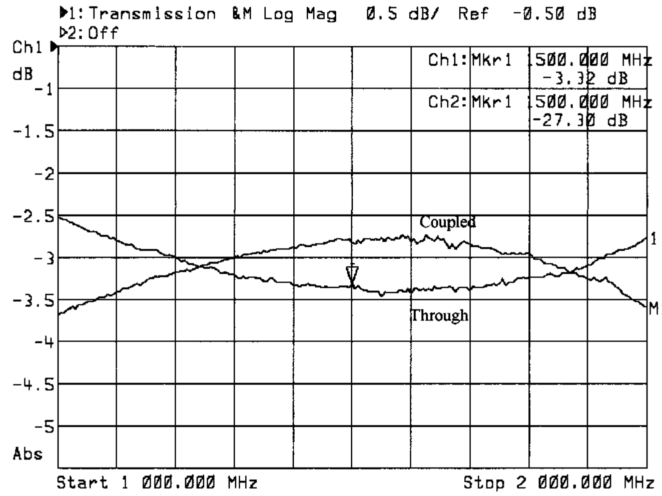

Fig. 10. Measured through and coupled response of the 1–2-GHz coupler.

IV. Measured Results

A. Octave-Band Circuit

It was decided that the lumped capacitors should be implemented as fixed components. Although the exact value could be arrived at using a variable component, for practical applications, it is undesirable to have to tune every coupler individually. It was found, in practice, that a lower value than predicted was required for optimum input return loss and isolation; the experimental value was 1.5 pF. Measured through and coupling response is shown in Fig. 10. The lower crossover occurs at 1.24 GHz, and the upper at 1.88 GHz, with maximum coupling at 1.6 GHz. The plot reveals the loss of the circuit. Ideally, the two traces should cross over at 3.01 dB. At 1.88 GHz, an excess loss of 0.2 dB is seen. The maximum imbalance occurs at 1 GHz, indicating insufficient coupling at this frequency. However, midband, it still achieves a value of 0.6 dB, as is predicted in the theoretical response of Fig. 5. It might be thought that the lower than calculated optimum capacitor value would be a cause of the insufficient coupling at 1 GHz. However, it has been found previously [8] that stray effects require a smaller than calculated capacitor value in microwave lumped-element hybrids.

Fig. 11 shows a plot of the measured isolation and input return loss. In the ideal circuit, these two quantities should be perfect. In practice, a worst-case value of 26.5 dB was measured for the isolation and 31 dB for the return loss. The coaxial-to-microstrip launch method and system accuracy add a significant measurement error to these numerical values, but the results are still excellent.

Fig. 12 shows a plot of the coupled to through ports relative phase. In the ideal circuit, this will be exactly 90° at all frequencies. In practice, the circuit deviated by less than 1° over the entire 1–2-GHz band.

Fig. 13 shows a plot of the loss to the isolation port, and input return loss with the through and coupled ports left open circuit. This measurement provides a figure-of-merit for the coupler. The theoretical lossless circuit of Fig. 2 exhibits a very small loss, which, at the maximum imbalance points, amounts to only

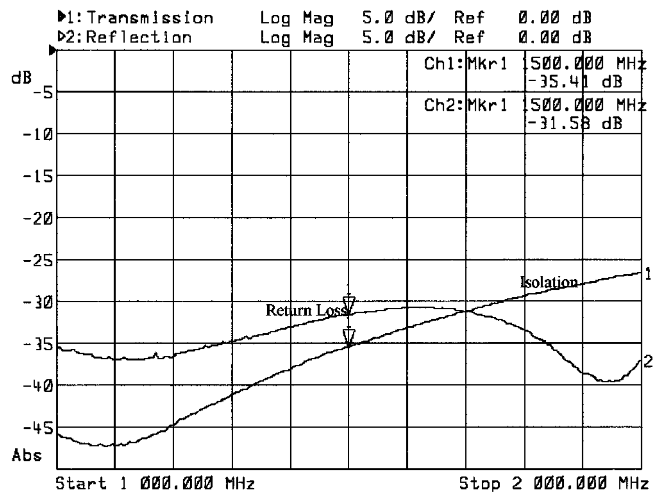

Fig. 11. Measured isolation and input return loss of the 1–2-GHz coupler.

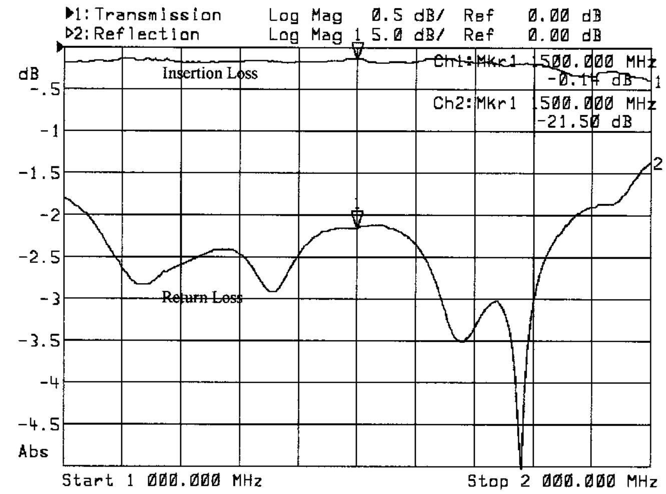

Fig. 13. Insertion loss to isolated port and input return loss with outputs O/C of the 1–2-GHz coupler.

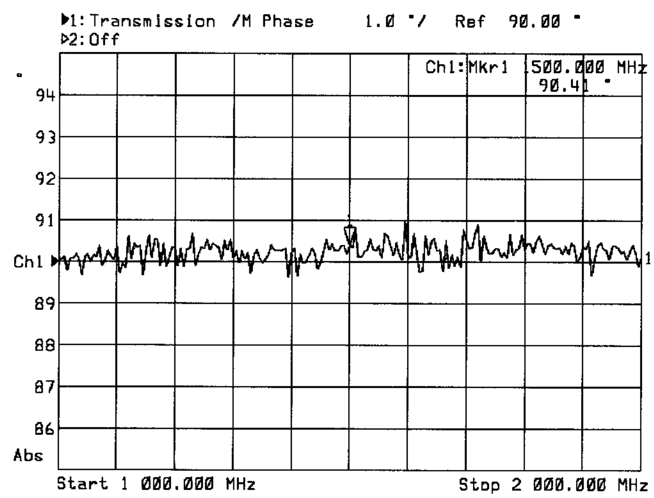

Fig. 12. Measured coupled to through relative phase of the 1–2-GHz coupler.

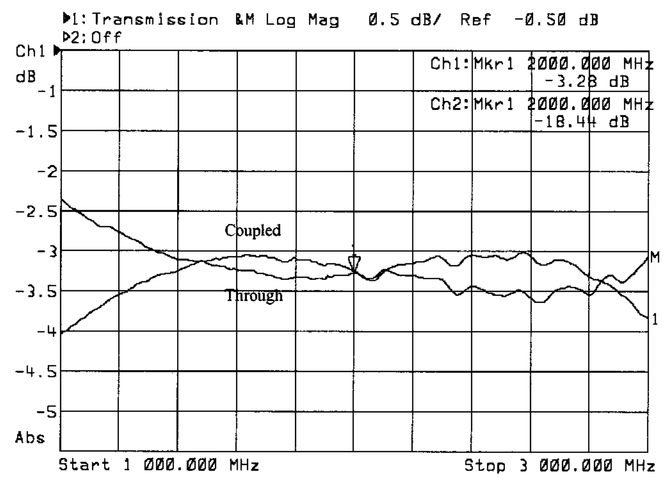

Fig. 14. Measured through and coupled response of the 1–2-GHz coupler.

0.024 dB, with a corresponding return loss of 22.6 dB. The practical circuit exhibits poorer performance than this, by virtue of resistive loss, discontinuities, and tolerances in element values. Despite the fact that the physical circuit showed maximum imbalance at the lower band edge, the worst-case loss was still at the upper band edge, where it achieves a value of 0.4 dB. The plot indicates that it would not be profitable to adjust the circuit for optimum amplitude balance, as the insertion loss is deteriorating toward 2 GHz and is likely to respond unfavorably to further compromise at the upper band edge.

B. Extended Octave-Band Circuit

Capacitor values for optimum isolation and input return loss were 2.4 pF for the center capacitor and 0.3 pF for the outer capacitors. Measured through and coupling response is shown in Fig. 14.

Crossovers occur at 1.5, 2.1, and 2.85 GHz. The middle and upper points are as expected, but the lower one has shifted from the predicted value of 1.155 GHz. This has contributed to the higher than designed amplitude imbalance at the lower band edge, amounting to 1.7 dB, and has also given an improved balance over the rest of the band. The predicted value was 0.64 dB, but above 1.2 GHz, it is never worse than 0.5 dB until right at the upper band edge. The upper crossover indicates the loss of the coupler, amounting to about 0.4 dB. This is more than the octave-band circuit, as might be expected, considering narrower lines were used, and the increase in frequency.

Fig. 15 shows a plot of the measured isolation and input return loss. The worst case isolation of 25 dB is acceptable in most applications. The worst-case input return loss of 21.5 dB is reasonable in view of the frequency range.

Fig. 16 shows a plot of the coupled to through ports relative phase. Taking into account the measurement system noise, it appears that the deviation from exact quadrature is 2° at most.

Fig. 17 shows a plot of the loss to isolated port and input return loss with the two outputs left open circuit. The large amplitude imbalance at the lower band edge has resulted in a poor return loss of only 12.6 dB, although the insertion loss has only slightly degraded. The insertion loss is again worse at the higher end of the band, reaching a value of 0.75 dB by 3 GHz. Toward

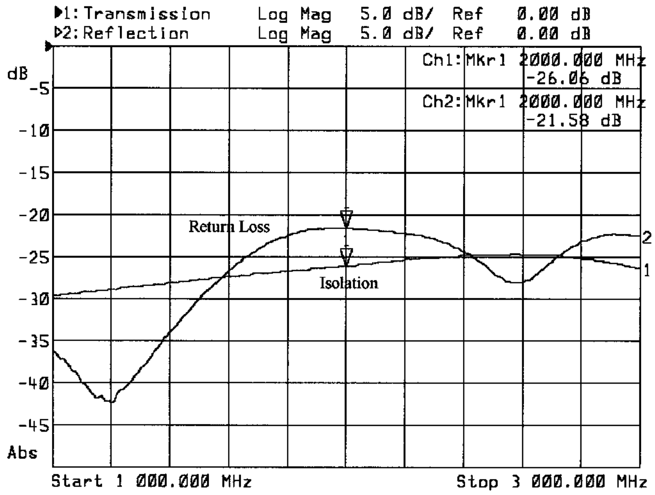

Fig. 15. Measured isolation and input return loss of the 1–3-GHz coupler.

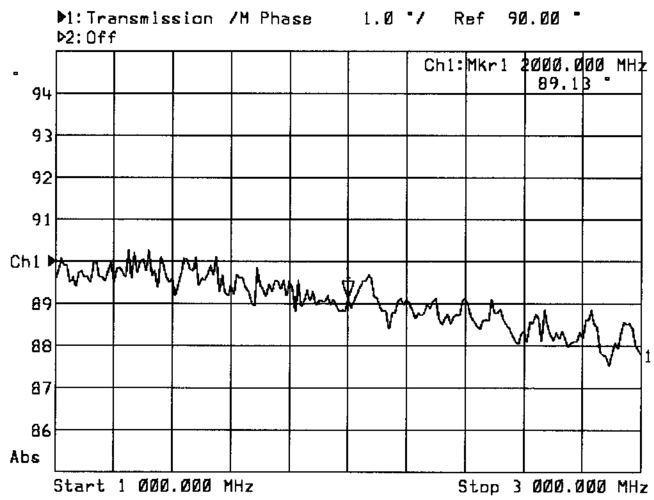

Fig. 16. Measured coupled-to-through relative phase of the 1–3-GHz coupler.

that frequency, the insertion loss is deteriorating rapidly. Once again, this would indicate that, in a practical use of this circuit, it would be better to leave it as it is, rather than optimize the coupling response.

V. Applications

The circuit for the octave-band coupler is applicable to bandwidths from about 20% to a little above an octave. Below this bandwidth range, the narrow-band lumped-element coupler designs of [8] will provide adequate performance in most applications. Above this range, the extended octave-band coupler will be necessary for acceptable amplitude balance, and will itself be adequate up to bandwidths between one-and-a-half and two octaves. Both of the coupler circuits are suitable where low dielectric-constant substrates are used, such as glass-filled PTFE and the newer thermoset resin-based materials. The use of higher dielectric materials, such as alumina, may give a difficulty in the extended octave circuit, in implementing the interconnecting lines over the central cavity, owing to their shorter physical length. However, when these materials are used, the circuits are often processed using thin-film technology, in which the existing Lange coupler designs are easily implemented.

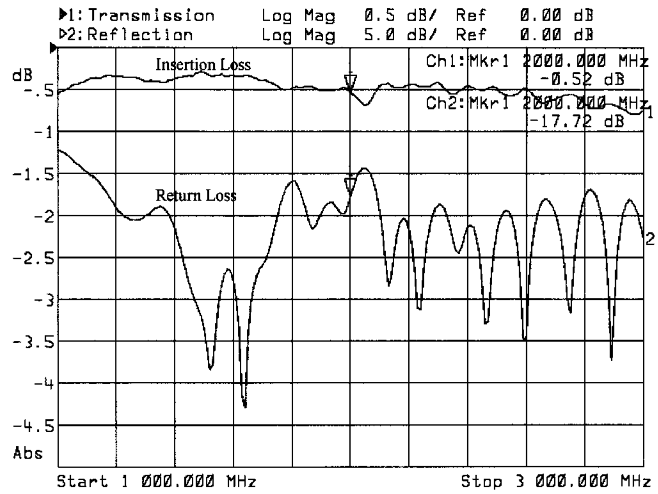

Fig. 17. Insertion loss to isolated port and return loss with outputs O/C of the 1–3-GHz coupler.

The new designs are particularly suitable for balanced amplifier stages, requiring quadrature hybrids on the input and output in their design. For this application, a carrier material is frequently available as a necessity for mounting high-power devices, into which the cavities can be machined. The new designs will be a ready alternative to custom coaxial or stripline hybrids, and can be designed for the particular frequency range. They will be especially useful above frequencies of around 4 GHz, where the existing transmission-line solutions give difficulties in implementation.

The power ratings of the new circuits will depend on the handling of the microstrip lines. For the particular examples given, this will be in the region of several tens of watts, though the 1–2-GHz coupler will give the better power handling, owing to its lower loss. The exact value will depend on the allowable conductor temperature rise, and operating temperature range. The power handling can be expected to be a little lower in the coupler region itself, compared with the rest of the microwave circuit, as the thermal path over the cavity will not be as good as the metal-backed areas.

The new designs are space efficient. This can be appreciated from the electrical length of the interconnecting lines. For the octave-band coupler, this amounted to only 19.5° at the center frequency. Although the remaining cavity portions will extend the physical length further, the complete assembly will still be significantly shorter than a quarter-wave coupler on the same substrate. The total physical length for the octave-band coupler was 24.5 mm, which is the length of a 71° line at the center frequency. For the extended octave-band coupler, the sum of connecting line lengths is 70.8° at the center frequency. As the end, lumped-element sections are only small structures, the total physical length is only slightly longer than what is necessary for the lines. This will still be less than a quarter-wave coupler.

As the new circuits are proposed to be replacements for existing design solutions, it would be appropriate to make comparisons of the performance.

It would be premature to make a full comparison with the Lange coupler. This circuit element comes into its own at higher microwave frequencies than those which have been considered above. The Lange coupler is most suitable for alumina substrates, which pose a difficulty for the new circuits, in that the connecting lines become physically shorter, leaving less space available for cavities. It is expected that the new designs will be preferable for low dielectric substrates, where standard PCB manufacturing technology is used, provided the low loss, high isolation, and good return loss can be maintained to higher frequencies.

The performance of the new circuits is within the specifications quoted for coaxial coupled lines, with the exception of power rating, which is greater for the coaxial line. It should be noted, however, that in a real circuit, the power will be limited by the microstrip lines anyway, thus, the full capability of the coaxial line will never be utilized. The new circuits will be an advantage above frequencies of around 3 GHz, where the coaxial line gives a poor microstrip transition.

The performance of the octave-band circuit exceeds that of surface mount hybrids in insertion loss, isolation, and return loss. Only the power rating of the surface mount hybrid is better, but again, this will not be utilized for reasons of microstrip line limitation. No comparison can be made with the extended octave-band circuit, as no commercial products are available for this bandwidth. Once again, the new circuits offer better prospects for operation at higher microwave frequencies, where surface-mount hybrids are not available.

If the octave-band circuit is compared with drop-in hybrids, it is seen that the specification is very similar. The drop-in hybrid has greater power handling, which again cannot be utilized in a microstrip circuit. The frequency limitation for the drop-in hybrid is similar to the surface-mount hybrid.

VI. Conclusions

The existing designs for wide-band 3-dB quadrature couplers, though unsuitable for microwave operation, can be adapted to give circuits that give good microwave performance.

Two designs have been presented: the first covering a frequency range of 1–2 GHz shows an excess loss of 0.2 dB, an input return loss of 31 dB, an isolation of 26 dB, and phase error from quadrature of less than 1°. With a maximum amplitude imbalance of 1.2 dB, it gives comparable performance to a single-section quarter-wavelength coupler.

The second covers a frequency range of 1–3 GHz, and shows an excess loss of 0.4 dB, an input return loss of 21.5 dB, an isolation of 25 dB, and phase error from quadrature of less than 2°. With a maximum amplitude imbalance of 1.7 dB, it gives performance in between a single- and three-section coupler. The designs are easy to implement and do not require high-precision technology. They are particularly appropriate for custom design problems, with significant power rating, such as wide-band microwave amplifiers.

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D. P. Andrews, photograph and biography not available at time of publication.

C. S. Aitchison (A'87–SM'90–F'94), photograph and biography not available at time of publication.