whence

$$|G''| \leqslant \sum_{n=2}^{\infty} a_n < \frac{e^{-(\pi/a)|z_{v'}-z'|\beta_2}}{\beta_2} \tag{B13}$$

as

$$\frac{e^{-(\pi/a)|z_{v'}-z'|\beta_{N+1}}}{\beta_{N+1}} \to 0 \text{ as } N \to \infty.$$

The second inequality in (B13) can also be deduced from Fig. 12. It follows from (B12) that G'' converges exponentially. Furthermore, since

$$a_n \leqslant b_n = \frac{1}{\beta_n} - \frac{1}{n}, \qquad n = 2,3,\cdots \tag{B14}$$

$$|G''| \leqslant \sum_{n=2}^{\infty} a_n \leqslant \sum_{n=2} b_n < \frac{1}{\beta_2} \tag{B15}$$

which can be proved using similar procedure, and is also evident from Fig. 12, G'' does converge uniformly for all (x', z').

Acknowledgment

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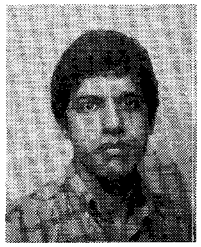

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Composite Coupler Design

THOMAS C. CHOINSKI

***Abstract* —Unequal power splitters and combiners are generally limited by the line widths which can be practically synthesized in a given transmission medium. This practical limitation on the ratio of unequal power division can be extended by incorporating the same types of couplers into a composite design.**

The general composite design approach outlined in this paper uses three couplers (three terminal couplers) to generate a new three-terminal circuit. The design equations are derived for the composite approach and summarized in graphic form.

The feasibility of the composite design approach is demonstrated by the construction of a 5.76-dB differential coupler using internally series-terminated Wilkinson couplers. The circuit was designed, analyzed via computer, and finally built and tested. The results from the composite design are compared to that of a single Wilkinson coupler design.

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I. Introduction

DIRECT THREE-TERMINAL couplers have found usefulness in many component applications including switches, phase shifters, amplifiers, and feed networks for antenna arrays. The networks may take the form of branch-line hybrids, Wilkinson couplers, rat-race magic tees, or simple tee structures which may incorporate an internally terminated fourth port. In any case, their primary function is to split and combine power in either an equal or weighted fashion.

The design of unequal three-terminal couplers which maintain equal output phase characteristics have been well defined in previously published literature. Network simplicity usually allows an adaption to familiar transmission media such as stripline, microstrip, or coax. However, limitations arise in their practical implementation. One of these limitations is the range of practical characteristic impedances which can be synthesized in applicable transmission-line structures. This restriction places a boundary on the differential coupling capability of a single unequal coupler.

A composite design approach is one method which can be used to extend the ability of a coupler to split power unequally. Previous work has been performed indicating the usefulness of cascaded couplers to obtain tight coupling values [7], [9]. The method outlined in this paper is used to achieve loose coupling. The technique consists of a combination of three-terminal splitters and combiners arranged to form a resultant three-terminal network with phase-matched outputs. The composite network yields higher output power differentials than any single integral coupler design given the same practical restraints on impedance formation.

The following discussion will develop the necessary criteria for a class of composite designs. Design equations will be summarized in graphic form to illustrate design trends. Finally, the general solution will be applied and validated using Wilkinson couplers with internal series terminations.

II. Design Principles

The composite design method initially involves the division of the input signal by two tiers of three-terminal power dividers, yielding three intermediate output terminals as shown in Fig. 1. Two of these intermediate output terminals are recombined via a third coupler. The addition of these two signals increases the power to one leg of the network, while the signal to the second output has been reduced by traveling through two dividing networks which are in series. Two additional phase-shifting components are required to maintain the phase balance of each leg.

The derivation of the design equations for this circuit can be completed using scattering matrices and an interconnection matrix describing the network topology of the system [4]. Initially, the forward transmission coefficients are derived. These coefficients are used to describe the total coupling differential for the composite network in terms of the coupling differentials of the individual couplers. Next, the reverse transmission coefficients are derived. The necessary requirements for reciprocity are defined.

The operation of this network can be described in terms of the individual scattering matrices of each coupler. The general form of the matrix equation is as follows:

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \\ b_4 \\ b_5 \\ b_6 \\ b_7 \\ b_8 \\ b_9 \\ b_{10} \\ b_{11} \\ b_{12} \\ b_{13} \end{bmatrix} = \begin{bmatrix} S_{11}^A & S_{12}^A & S_{13}^A & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ S_{21}^A & S_{22}^A & S_{23}^A & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ S_{31}^A & S_{32}^A & S_{33}^A & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & S_{11}^B & S_{12}^B & S_{13}^B & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & S_{21}^B & S_{22}^B & S_{23}^B & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & S_{31}^B & S_{32}^B & S_{33}^B & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & S_{11}^C & S_{12}^C & S_{13}^C & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & S_{21}^C & S_{22}^C & S_{23}^C & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & S_{31}^C & S_{32}^C & S_{33}^C & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & S_{11}^D & S_{12}^D & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & S_{21}^D & S_{22}^D & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & S_{11}^E & S_{12}^E \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & S_{21}^E & S_{22}^E \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ a_3 \\ a_4 \\ a_5 \\ a_6 \\ a_7 \\ a_8 \\ a_9 \\ a_9 \\ a_{10} \\ a_{11} \\ a_{12} \\ a_{13} \end{bmatrix}. \quad (1)$$

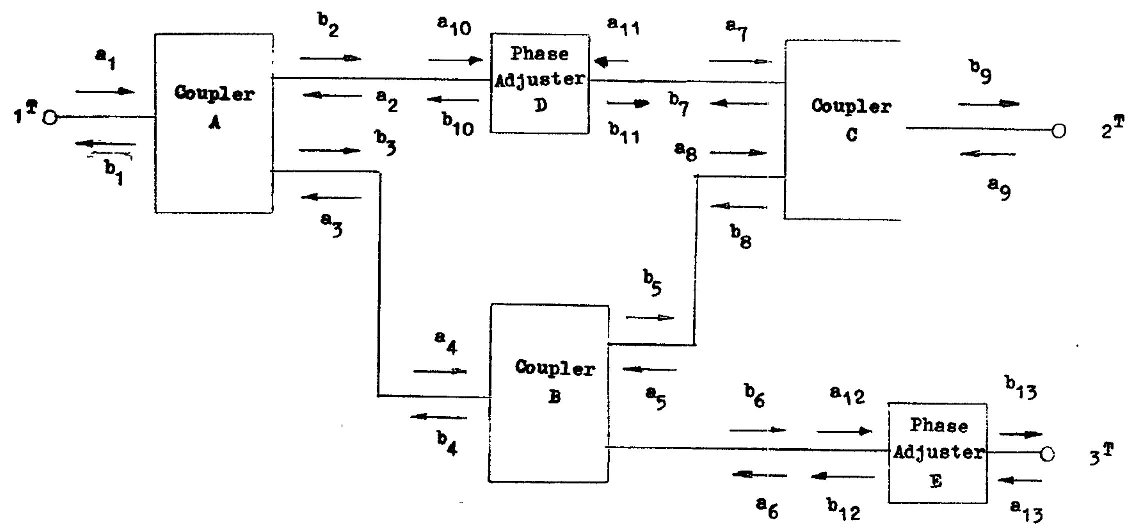


Fig. 1. General composite coupling network.

Likewise, the topology of the overall circuit can be represented by its interconnection matrix. The interconnection matrix for the network in Fig. 1 is

$$\begin{bmatrix} b_1\\ b_2\\ b_3\\ b_4\\ b_5\\ b_6\\ b_7\\ b_8\\ b_9\\ b_{10}\\ b_{11}\\ b_{12}\\ b_{13} \end{bmatrix} = \begin{bmatrix} 0&0&0&0&0&0&0&0&0&0&0&0&0\\ 0&0&0&0&0&0&0&0&0&1&0&0&0\\ 0&0&0&1&0&0&0&0&0&0&0&0&0\\ 0&0&1&0&0&0&0&0&0&0&0&0&0\\ 0&0&0&0&0&0&0&1&0&0&0&0&0\\ 0&0&0&0&0&0&0&0&0&0&0&1&0\\ 0&0&0&0&0&0&0&0&0&0&1&0&0\\ 0&0&0&0&1&0&0&0&0&0&0&0&0\\ 0&0&0&0&0&0&0&0&0&0&0&0&0\\ 0&1&0&0&0&0&0&0&0&0&0&0&0\\ 0&0&0&0&0&0&1&0&0&0&0&0&0\\ 0&0&0&0&0&1&0&0&0&0&0&0&0\\ 0&0&0&0&0&0&0&0&0&0&0&0&0 \end{bmatrix} \begin{bmatrix} a_1\\ a_2\\ a_3\\ a_4\\ a_5\\ a_6\\ a_7\\ a_8\\ a_9\\ a_{10}\\ a_{11}\\ a_{12}\\ a_{13} \end{bmatrix}. \tag{2}$$

The forward transmission coefficients S_{31}^T and S_{21}^T for the entire network can be derived if the scattering matrix can be reduced, by assuming each individual three-port network is matched and maintains isolation between output ports, i.e., $S_{11}^{A,B,C,D,E} = S_{22}^{A,B,C,D,E,} = S_{33}^{A,B,C} = 0$ and $S_{23}^{A,B} = S_{32}^{A,B} = S_{12}^C = S_{21}^C = 0$. These conditions assure the absence of any traveling waves in the reverse direction when the input port to coupler A is fed. The following equations and reduced scattering matrix can be concluded for this case:

$$b_1 = b_4 = b_7 = b_8 = b_{10} = b_{12} = 0 \tag{3}$$

$$a_2 = a_3 = a_5 = a_6 = a_9 = a_{11} = a_{13} = 0 \tag{4}$$

$$\begin{bmatrix} b_2\\ b_3\\ b_5\\ b_6\\ b_9\\ b_{11}\\ b_{13} \end{bmatrix} = \begin{bmatrix} S_{21}^A&0&0&0&0&0\\ S_{31}^A&0&0&0&0&0\\ 0&S_{21}^B&0&0&0&0\\ 0&S_{31}^B&0&0&0&0\\ 0&0&S_{31}^C&S_{32}^C&0&0\\ 0&0&0&0&S_{21}^D&0\\ 0&0&0&0&0&S_{21}^E \end{bmatrix} \begin{bmatrix} a_1\\ a_4\\ a_7\\ a_8\\ a_{10}\\ a_{12} \end{bmatrix}. \tag{5}$$

The resulting expressions for the forward transmission coefficients of the entire network are easily derived through the algebraic manipulation of the interconnection equations (2) and the reduced S matrix (5). These solutions are given for midband as follows:

$$S_{21}^{T}=\left.\frac{b_9}{a_1}\right|_{\substack{a_9=0\\a_6=0}}=S_{21}^{A}S_{31}^{C}S_{21}^{D}+S_{31}^{A}S_{21}^{B}S_{32}^{C} \tag{6}$$

$$S_{31}^{T}=\left.\frac{b_{13}}{a_1}\right|_{\substack{a_9=0\\a_6=0}}=S_{31}^{A}S_{31}^{B}S_{21}^{E}. \tag{7}$$

The equation for S_{21}^{T} can be further simplified through a judicious selection of S_{21}^{D}. A single length of lossless transmission line having a characteristic impedance equal to 50 Ω can be used to equalize the phase of both terms in (6), permitting them to add directly, while setting the magnitude of S_{21}^{D} equal to one. Likewise, an identical selection of S_{21}^{E} will equalize the transmission phases of S_{21}^{T} and S_{31}^{T} provided couplers A, B, and C have the same phase characteristics. Therefore, the magnitudes of S_{21}^{T} and S_{31}^{T} would be

$$|S_{21}^{T}|=|S_{21}^{A}\quad S_{31}^{C}|+|S_{31}^{A}\quad S_{21}^{B}\quad S_{32}^{C}| \tag{8}$$

$$|S_{31}^{T}|=|S_{31}^{A}\quad S_{31}^{B}|. \tag{9}$$

Given these expressions for the transmission coefficients, the coupling ratio of the output power from both ports of the composite circuit can be deduced and described in terms of the coupling ratios of the individual couplers as follows:

$$K_T=\frac{|S_{21}^{T}|^2}{|S_{31}^{T}|^2}=\frac{1-|S_{31}^{T}|^2}{|S_{31}^{T}|^2}=K_A+K_AK_B+K_B \tag{10}$$

where

$$K_{A,B}=\frac{|S_{21}^{A,B}|^2}{|S_{31}^{A,B}|^2} \tag{11}$$

$$|S_{21}^{T}|^2+|S_{31}^{T}|^2=1 \tag{12}$$

and

$$|S_{21}^{A,B}|^2+|S_{31}^{A,B}|^2=1. \tag{13}$$

Interestingly, the overall power output differential is only a function of the output power ratios of couplers A and B. Likewise, the necessary differential for coupler C will be subject to the values of K_A and K_B. The solution for K_C can be found by substituting (8) into (10) to obtain the equality

$$\left|\,|S_{21}^{A}\quad S_{31}^{C}|+|S_{31}^{A}\quad S_{21}^{B}\quad S_{32}^{C}|\,\right|^2=1-|S_{31}^{A}\quad S_{31}^{B}|^2. \tag{14}$$

Solving for K_C

$$K_C=K_A+\frac{K_A}{K_B} \tag{15}$$

where

$$K_C=\frac{|S_{31}^{C}|^2}{|S_{32}^{C}|^2} \tag{16}$$

and

$$|S_{31}^{C}|^2+|S_{32}^{C}|^2=1. \tag{17}$$

The composite coupler may also be used as a combiner by feeding the two output ports. In fact, if reciprocity is maintained, the circuit will combine two signals in the same weighted fashion that it would split a single input signal. Intuitively, we can expect to achieve overall reciprocity by the fact that each coupler is reciprocal. However, the requirements for this condition can be absolutely determined by solving for S_{13}^{T} and S_{12}^{T} using the same assumption that was used for S_{31}^{T} and S_{21}^{T}. Thus, for combiner operation, only reverse traveling waves will be produced and all forward traveling waves will be eliminated. This procedure will yield the following conditions and reduced scattering matrix:

$$b_2=b_3=b_5=b_6=b_9=b_{11}=b_{13}=0 \tag{18}$$

$$a_1=a_4=a_7=a_8=a_{10}=a_{12}=0 \tag{19}$$

$$\begin{bmatrix} b_1\\ b_4\\ b_7\\ b_8\\ b_{10}\\ b_{12}\end{bmatrix}=\begin{bmatrix} S_{12}^{A} & S_{13}^{A} & 0 & 0 & 0 & 0 & 0\\ 0 & 0 & S_{12}^{B} & S_{13}^{B} & 0 & 0 & 0\\ 0 & 0 & 0 & 0 & S_{13}^{C} & 0 & 0\\ 0 & 0 & 0 & 0 & S_{23}^{C} & 0 & 0\\ 0 & 0 & 0 & 0 & 0 & S_{12}^{D} & 0\\ 0 & 0 & 0 & 0 & 0 & 0 & S_{12}^{E}\end{bmatrix}\begin{bmatrix} a_2\\ a_3\\ a_5\\ a_6\\ a_{11}\\ a_{13}\end{bmatrix}. \tag{20}$$

Combining these sets of equations with the interconnection matrix determined S_{12}^{T} and S_{13}^{T} as follows:

$$S_{12}^{T}=\left.\frac{b_1}{a_9}\right|_{\substack{a_6=0\\a_9=0}}=S_{12}^{A}S_{13}^{C}S_{12}^{D}+S_{13}^{A}S_{12}^{B}S_{23}^{C} \tag{21}$$

$$S_{13}^{T}=\left.\frac{b_1}{a_{13}}\right|_{\substack{a_1=0\\a_9=0}}=S_{13}^{A}S_{13}^{B}S_{12}^{E}. \tag{22}$$

A quick inspection of (6), (7), (21), and (22) will show that reciprocity will be preserved if $S_{12}^{A,B,D,E}=S_{21}^{A,B,D,E}$, $S_{13}^{A,B}=S_{31}^{A,B}$, $S_{23}^{C}=S_{32}^{C}$, and $S_{13}^{C}=S_{31}^{C}$. Thus, each individual coupling network must be matched, reciprocal, and possess two isolated output ports in order for the composite circuit to maintain reciprocity. The only possible way to achieve these individual coupler requirements is to use designs which incorporate some kind of lossy element as outlined by Carlin and Giordano [1]. The stipulation of a lossy three port does not hinder the operation of the composite coupler. In fact, it is the only way to achieve the desired performance. In other words, discretion must be exercised in the choice of the type of individual coupler which is incorporated into the design.

The aforementioned design equations leave three degrees of freedom K_A, K_B, and K_C for a particular value of K_T. Therefore, additional restrictions must be placed on the design to arrive at a solution. These requirements can be concluded arbitrarily. Two methods will be developed in this paper.

The first method involves picking a desired value for K_C and using subsequent values obtained for K_A and K_B. For instance, K_C may be picked to minimize the coupling ratio

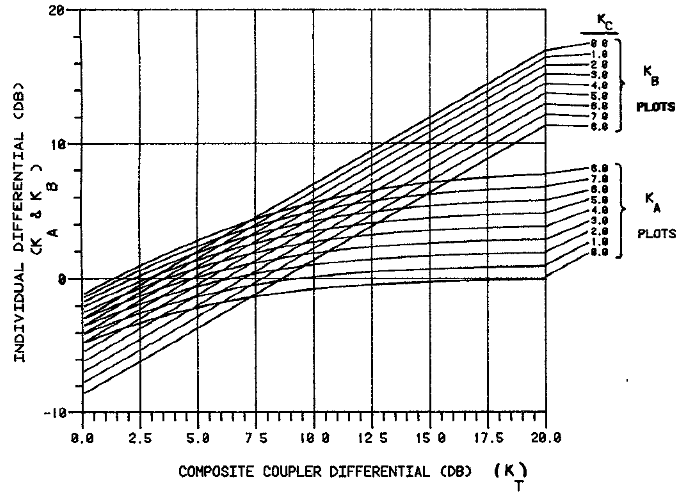


Fig. 2. Composite couplers.

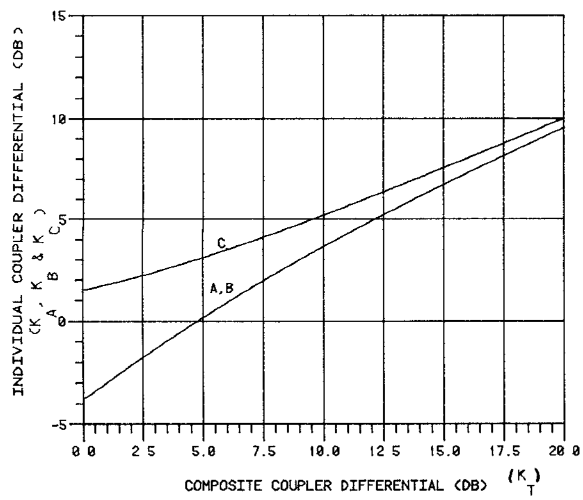


Fig. 3. Composite couplers ($K_A = K_B$).

of coupler C, i.e., $K_C = 1$. Once the value has been decided upon, the remaining couplers can be derived from (10) and (15) as follows:

$$K_A = \frac{K_C K_T}{K_C + K_T + 1} \tag{23}$$

$$K_B = \frac{K_T}{K_C + 1}. \tag{24}$$

Plots of these solutions are shown in Fig. 2 for various values of K_C.

An alternative method is to equalize the coupling ratios of couplers A and B. When this approach is pursued, the following values are obtained:

$$K_A = K_B = \sqrt{(1+K_T)} - 1 \tag{25}$$

$$K_C = \sqrt{(1+K_T)}. \tag{26}$$

Fig. 3 shows a graph of the results from this solution. It is interesting to note that for this case the largest integral coupling differential required to synthesize the composite coupler is approximately one half the total desired unequal power split (in decibels). The total unequal splitting capability has essentially been squared.

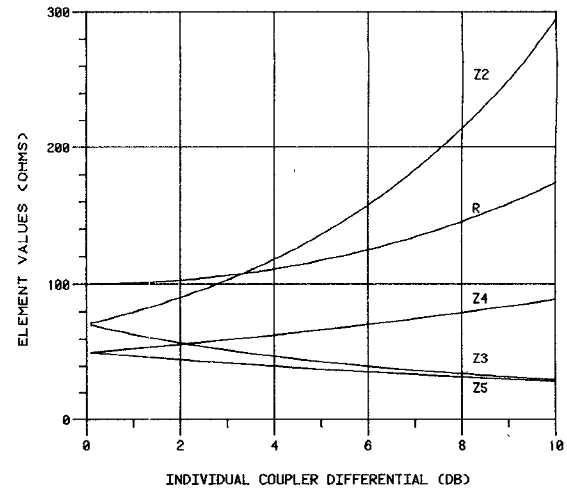


Fig. 4. Unequal Wilkinson couplers.

TABLE I
TWO-STAGE WILKINSON COUPLERS ($Z_0 = 50.0\ \Omega$)

COUPLING DIFFERENTIAL (dB)	Z_2	Z_3	Z_4	Z_5	R
0.73 (A and B)	65.13	77.05	47.94	52.15	100.35
3.39 (C)	49.67	108.42	41.14	60.77	107.71
5.76 (single)	40.37	152.09	35.89	69.66	122.81

III. WILKINSON APPLICATION

An appropriate implementation of the prescribed design procedure can be completed using Wilkinson Couplers. These devices will be capable of providing the required power division while maintaining the necessary reciprocal characteristics. The key to the fulfillment of this criteria is the internal series termination of the Wilkinson's fourth port. The isolation resistor dissipates energy from unequal output reflections while maintaining good input and output match conditions.

The design of series terminated, three-port in-line unequal split Wilkinson Couplers has been previously outlined in published literature [2], [5], [6], [10]. The resulting design equations for the two-section uncompensated case as outlined by Howe are shown as follows:

$$K = \sqrt{P_b/P_a} \tag{27}$$

$$Z_2 = Z_0\sqrt{K(1+K)^2} \tag{28}$$

$$Z_3 = Z\sqrt{(1+K^2)/K^3} \tag{29}$$

$$Z_4 = Z_0\sqrt{K} \tag{30}$$

$$Z_5 = Z_0/\sqrt{K} \tag{31}$$

$$R = \left[\frac{1+K^2}{K}\right]Z_0. \tag{32}$$

The component values versus the power division ratio can easily be plotted to determine practical boundaries for

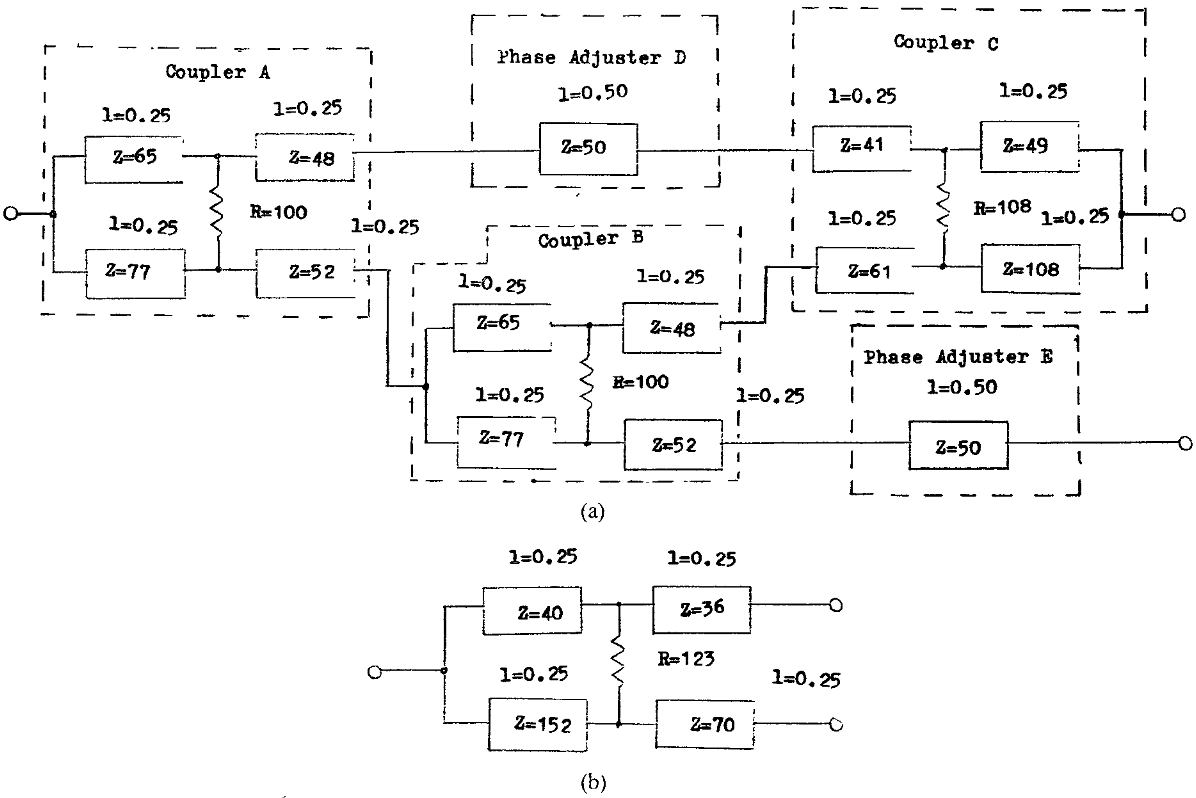


Fig. 5 5.76-db unequal coupler circuits. (a) Composite coupler. (b) Wilkinson coupler. All lengths are given in wavelengths. All impedances are given in ohms.

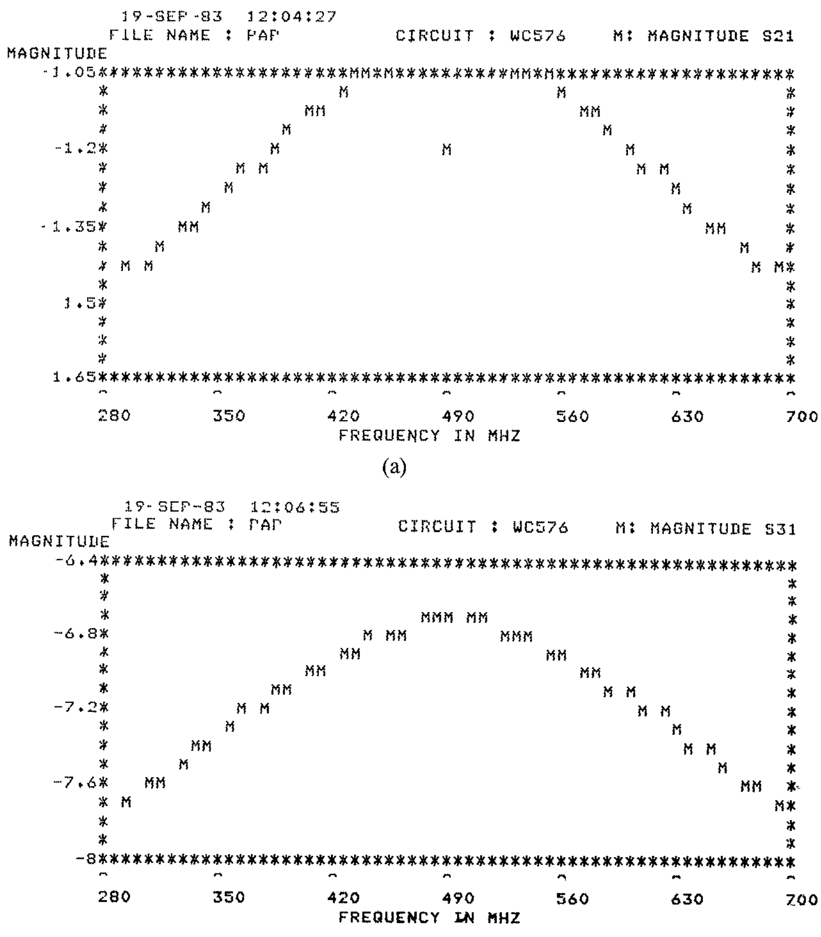
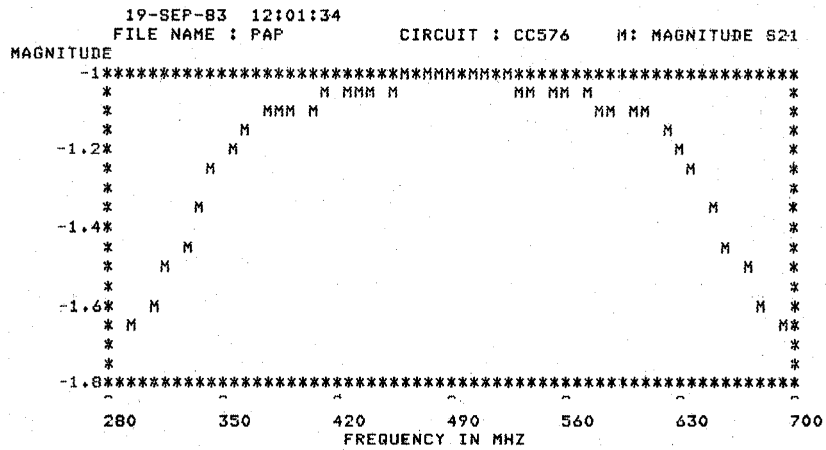
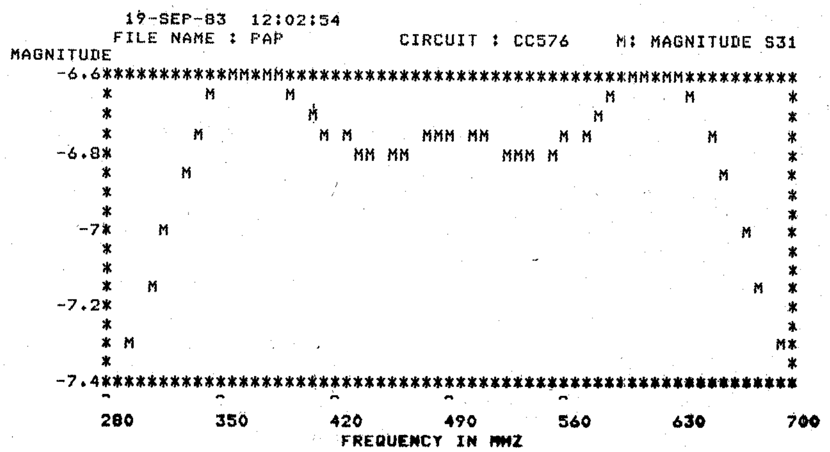


Fig. 6. S_{21} and S_{31} for 5.76-db Wilkinson coupler. (a) S_{21} versus frequency for unequal Wilkinson coupler. (b) S_{31} versus frequency for unequal Wilkinson coupler.



(a)



(b)

Fig. 7. S_{21} and S_{31} tor 5.76-db composite coupler. (a) S_{21} versus frequency for composite coupler. (b) S_{31} versus frequency for composite coupler.

coupler synthesis (see Fig. 4). In general, transmission-line structures allow practical construction of transmission lines with characteristic impedances ranging from 20 Ω to approximately 120 Ω. If we use this range as our guideline, there will be a boundary due to the value of Z_2 near a coupling differential of 4.0 dB. Therefore, any coupler requiring an output ratio greater than 4.0 dB will be difficult to fabricate due to the small physical width associated with transmission line Z_2. An alternate technique must be used to achieve higher differential coupling values.

IV. Design Example

A coupler has been designed using the composite coupler technique. The circuit required a coupling differential of 5.76 dB. The intended application is a corporate feed network for a low sidelobe UHF dipole array which requires output phase tracking and accurate coupling. For simplicity, the design has been completed allowing couplers A and B to be equalized as previously described. The required phase equalizers where fabricated as half wavelength 50-Ω transmission lines.

The component values of the integral Wilkinson Couplers are given in Table I for both the composite design and the single unequal split coupler (which is practically unrealizable). The circuit diagram for each can be seen in Fig. 5

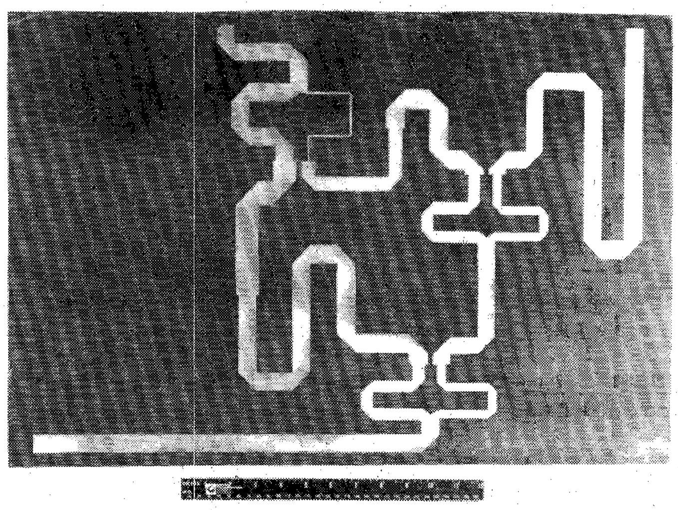


Fig. 8. Photograph of composite coupler.

The circuit complexity for both couplers demands a computer-aided analysis for each to determine performance. The analysis can be completed by using the techniques outlined in [4] or by using commercially available software such as SUPER COMPACT™. The latter was used in this case because of its easy access. A file was created describing each circuit, and the overall scattering parameters were obtained for both. The forward transmis-

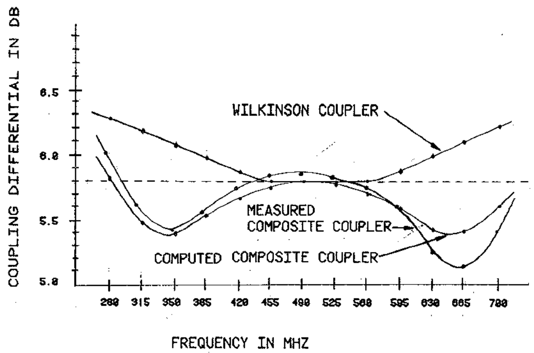


Fig. 9. Differential coupling versus frequency.

sion coefficients are shown versus frequency for the Wilkinson coupler of Fig. 5(b) in Fig. 6 and for the composite coupler of Fig. 5(a) in Fig. 7.

The 5.76-dB coupler was built and tested to verify the design. A suspended stripline structure was used. The ground plane spacing was 0.5 in, and the center conductor was etched on a 0.010-in woven teflon glass board. A photograph of the final circuit can be seen in Fig. 8.

Data was taken on the composite coupler using a network analyzer. The results are given in Fig. 9. The theoretical data obtained from the computer analysis of the Wilkinson and composite couplers are also presented for comparison. The composite coupler performed according to the computer analysis.

V. Conclusion

There are several different types of couplers which can be used to obtain an unequal power division of an input signal. Traditional couplers are satisfactory until the output power differentials become so high that the line widths of the transmission lines used in the designs become impractical to fabricate. When this situation occurs, another design procedure must be used.

One alternative design is the composite coupler. This approach uses several couplers to split and recombine the input signal to achieve greater unequal power output differentials. In some cases, the practical obtainable output coupling differentials are nearly doubled (in decibels).

The generalized performance characteristics depend on the type of coupler which is integrated into the composite design. In general, the power handling capability, output phase tracking, and VSWR will be similar to the individual characteristics of the couplers incorporated into the design.

The performance of the composite coupler is comparable to that of the series-terminated Wilkinson coupler when series-terminated Wilkinson couplers are used in the composite design. The only disadvantages are the additional resistors required and the increased size due to circuit topology. Nevertheless, the composite design approach does provide a viable alternative when high-output power differentials are required for a particular design application.

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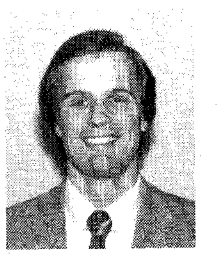

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