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Symmetrical Combiner Analysis Using *S*-Parameters

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***Abstract*—A general theory is developed to predict the potential efficiency (η) and input impedance (Z_{ic}) of symmetrical N-way combining networks in terms of scattering parameters. A simplified version of the theory, assuming perfect symmetry, is then implemented on a semiautomatic network analyzer (SANA) which is used to characterize 2-way and 16-way TM_{010} combining networks.**

These simplified theoretical assumptions have also been used to predict the degradation effects of power combiners when one or more sources fail. Results indicate that there is room for improvement if proper design techniques are applied.

I. INTRODUCTION

PRACTICAL REALIZATION of solid-state microwave transmitters are now feasible due to the continuing improvement of solid-state microwave power devices. In applications where power levels of individual microwave solid-state devices are insufficient to satisfy system requirements, it becomes necessary to efficiently combine many devices to reach the desired power levels. The purpose of this paper is to develop techniques to design, analyze, and characterize efficient solid-state power-combining networks, as well as to present some experimental verification of these techniques. In addition, the possibility of improving "graceful degradation" characteristics will be explored. In general, the theoretical portions of the approaches given here are applicable only to symmetrical power-combining networks.

In the past, solid-state power-combiner design has been implemented by integrating the device matching networks into the power-combining structure. With this approach, it is difficult to isolate problems to either device or combiner and it is also required that alignment to obtain maximum power be done experimentally. The approach presented here will be to separate the total combiner into individual modules. This will 1) simplify analysis, 2) allow alignment

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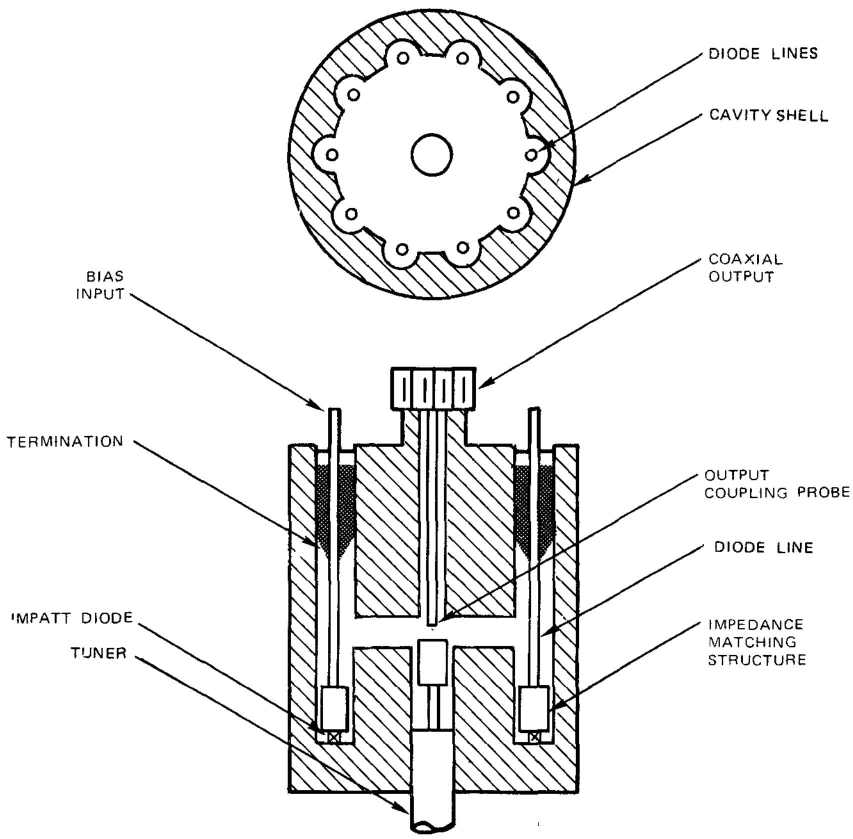


Fig. 1. Schematic drawing of a cylindrical cavity TM_{010} mode power combiner.

and characterization on a network analyzer for optimum efficiency, and 3) allow possible design modification(s) to improve the degradation characteristics that occur upon failure of individual solid-state devices.

A comparison of general combining techniques has recently been published [1] which separates summing networks into two basic types—those that combine the outputs of N devices in a single step (N-way combiners) and those that do not.

Since power is at a premium, the goal of combining large numbers of devices at the highest possible efficiencies makes the N-way summing model the best choice, as the N devices are combined directly in one step and do not have to pass through many stages with any attendant losses.

N-way summing networks can be subdivided further into resonant and nonresonant structures. When combining negative resistance devices, such as IMPATT diodes, the resonant structure has the advantage of eliminating spurious modes of oscillation while giving up the broad bandwidth of the nonresonant approach. Fortunately, as the number of devices to be combined (N) increase, efficiency and bandwidth of the resonant N-way structure improves [2], which is the opposite of serial and corporate structures.

In this report, a TM_{010} N-way summing network design (Fig. 1) will be used for experimental verification of the characterization techniques.

All theoretical equations will be presented in terms of the scattering parameters, since they are readily measured at microwave frequencies, thus eliminating any complicated parameter conversions.

II. Theory

A. The Potential Efficiency of a Symmetrical N-Way Summing Network in Terms of Its Unmatched Scattering Parameters

A block diagram of a symmetrical combiner is shown in Fig. 2 with its attendant matching circuits; it is redrawn in a more general way in Fig. 3. The power output is proportional to $|E_{\Sigma}^{-}|^2$, while the power input is proportional to $\Sigma_{j=1}^{N}|E_j|^2$, with the same proportionality constant if the fields are properly normalized to the characteristic impedances—which will be assumed to be the case. The efficiency is thus

$$\eta=|E_{\Sigma}^{-}|^2/N|E_j|^2 \tag{1}$$

where it is assumed that all of the sources are identical ($E_i=E_j$).

The input and output signals of the unmatched network are related by the $(N+1)\times(N+1)$ scattering matrix

$$(E^-)=(S)(E^+) \tag{2}$$

where (E^-) and (E^+) are $(N+1)$ column matrices. By symmetry $S_{ii}=S_{jj}$; $S_{ij}=S_{i\pm k,j\pm k}$ and $S_{i\Sigma}=S_{j\Sigma}$ where i, j, and k are running indexes from 1 to N for properly normalized ports. Reciprocity is assumed and requires that

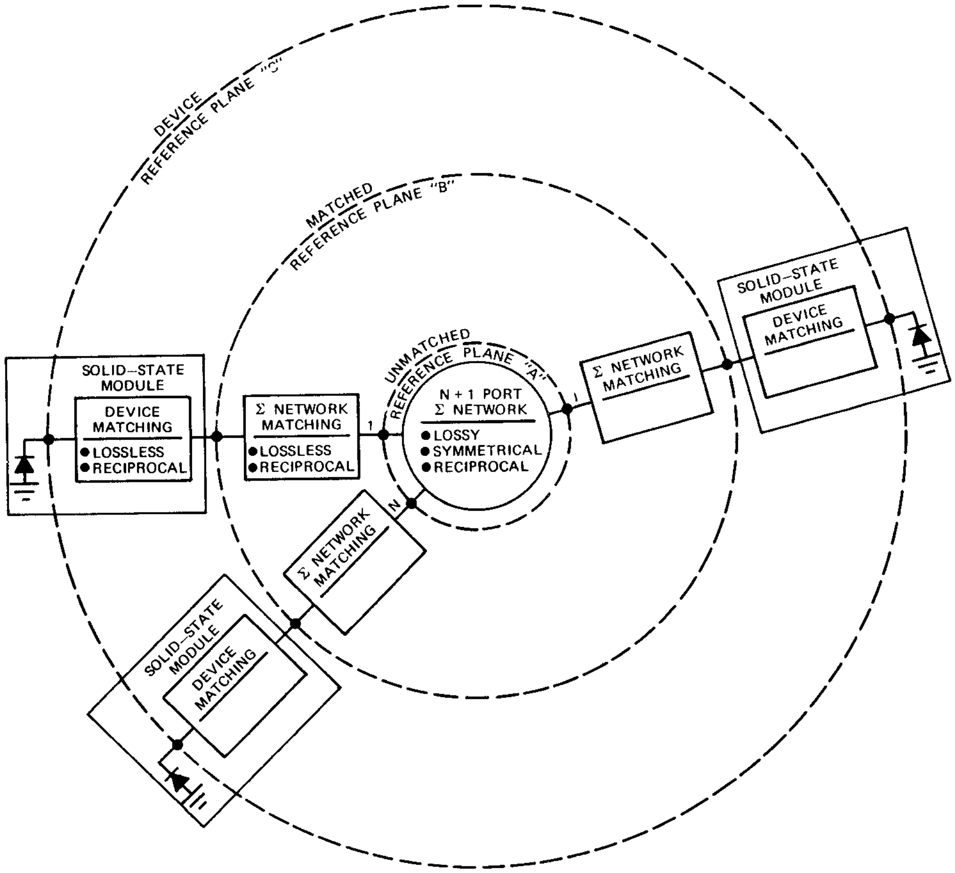


Fig. 2. Power combiner block diagram.

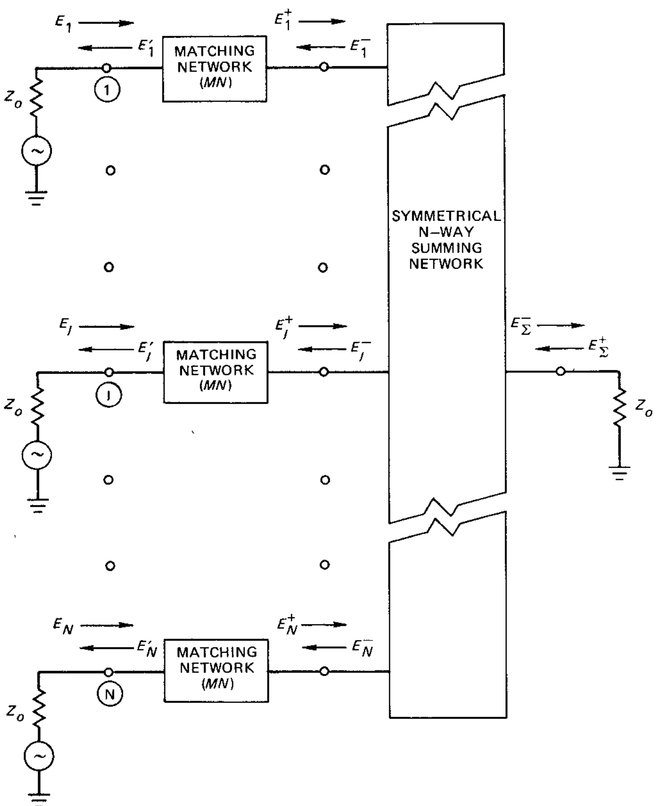


Fig. 3. Block diagram of the matched summing network.

$S_{ij} = S_{ji}$. It will also be assumed that the output port is properly terminated, i.e., $E_\Sigma^+ = 0$.

It follows that

$$E_\Sigma^- = NS_{i\Sigma}E_j^+ \tag{3}$$

which may be substituted into (1). For further reduction, the signals E_j^+ are needed in terms of E_j. Under the conditions specified ($E_i = E_j$ for all device ports)

$$E_i^- = S_L E_i^+ \tag{4}$$

where

$$S_L = \sum_{j=1}^{N} S_{ij}.$$

Let the matching networks (MN) in Fig. 3 be lossless and reciprocal, with a 2×2 scattering matrix (a). The unitary condition then requires

$$|a_{11}| = |a_{22}| \tag{5a}$$

$$\phi_{11} = 2\phi_{12} - \phi_{22} \pm \pi \tag{5b}$$

$$a_{12}^2 - a_{11}a_{22} = \exp j(2\phi_{12}) \tag{5c}$$

where ϕ_{ij} is the phase associated with a_{ij}.

Symmetry requires that all matching networks be identical, leading from (4) to

$$E_j^+ = a_{12}E_j/(1 - a_{22}S_L). \tag{6}$$

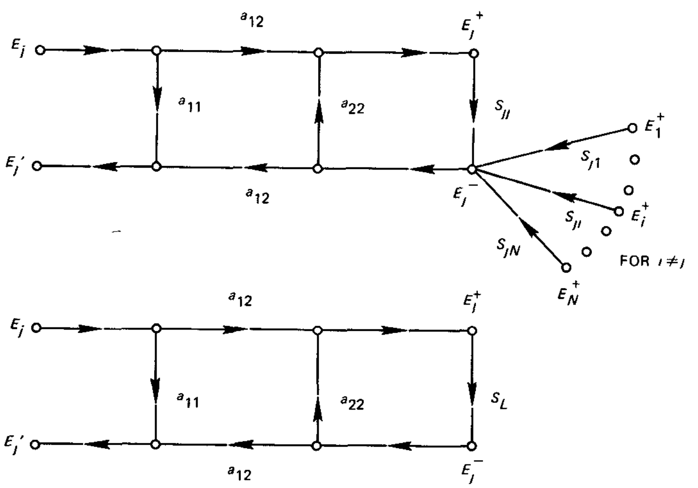


Fig. 4. Signal flow graph of the matching network.

We now make the assumption that the sources in Fig. 3 are matched, i.e.,

$$a_{11}E_i + a_{12}E_i^- = 0 \tag{7}$$

which follows directly from $E_i' = 0$.

Using (4) and (6), the following is readily derived from (7):

$$a_{11}(1 - a_{22}S_L) + a_{12}^2 S_L = 0. \tag{8}$$

With the foregoing, the efficiency can be written as

$$\eta = N|S_{i\Sigma}|^2 |a_{12}|^2 / |1 - a_{22}S_L|^2. \tag{9}$$

In order to obtain the efficiency solely in terms of the S-parameters of the combiner network, (8) can be solved with the aid of (5) to give

$$a_{22} = S_L^* \tag{10}$$

which leads to

$$\eta = N|S_{i\Sigma}|^2 / (1 - |S_L|^2). \tag{11}$$

When the scattering parameters of the unmatched combiner are measured, there will be a nonzero reflection coefficient at each input port and an input impedance given by

$$Z_{ic} = Z_0(1 + S_L)/(1 - S_L). \tag{12}$$

It is important to note that the inputs (E_j) to the matched cavity combiner are matched only under the condition of N identical sources.

It is instructive to look at a signal flow graph of an input port, as shown in Fig. 4. It is seen that S_L represents a composite scattering coefficient, and it is fairly simple to show that $|S_L| \leq 1$, as required for power to be put into the combining network rather than extracted at the input ports.

B. Determining the Scattering Parameters of a Symmetrical N-Way Summing Network by Measurements with N − 2 Ports Shorted

The usefulness of the theory just presented is largely determined by our ability to measure the scattering parameters. This is not always easy, as reference to Fig. 1 shows, since the device ports are not usually readily accessible for attaching the necessary connectors and matched loads. It may not only be expensive and time-consuming to attach them but, as N becomes large, there may not be adequate physical space to mount the connectors.

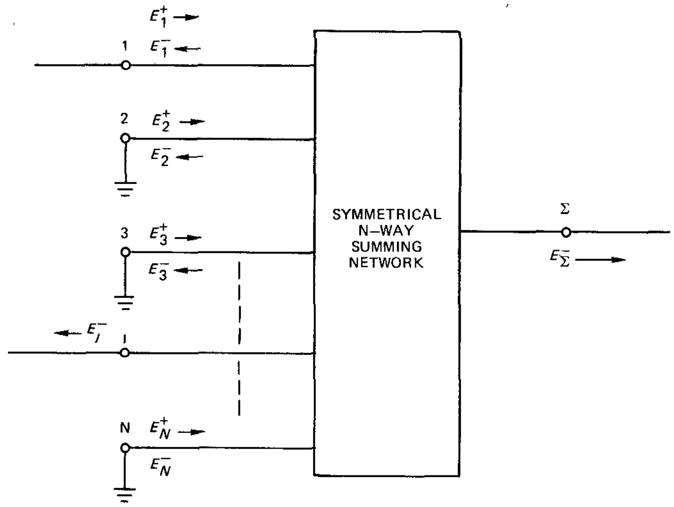


Fig. 5. Block diagram for N-way summing network with $N-2$ ports short-circuited.

This section of the paper is devoted to an alternate approach in which only two input ports, in addition to the output or sum port, are used, and the remaining ports are shorted—an operation that is relatively simple to carry out. The method is equally applicable at any reference plane in Fig. 2—the combiner itself (reference plane "A") or combiner plus matching (reference planes "B" or "C")—depending on where the shorts are placed; however, the shorted reference plane must be placed so as not to reflect a short circuit at the summing point.

Consider Fig. 5 which represents the combining network with all but the first and jth port shorted. The first and jth ports are terminated in a matched load. The measured scattering parameters are, therefore, those of a three-port network and are given by

$$b_{11} = E_1^- / E_1^+ |_{E_\Sigma^+ = E_j^+ = 0} \tag{13a}$$

$$b_{j1} = E_j^- / E_1^+ |_{E_\Sigma^+ = E_j^+ = 0} \tag{13b}$$

$$b_{\Sigma 1} = E_\Sigma^- / E_1^+ |_{E_\Sigma^+ = E_j^+ = 0} \tag{13c}$$

$$b_{\Sigma\Sigma} = E_\Sigma^- / E_\Sigma^+ |_{E_1^+ = E_j^+ = 0}. \tag{13d}$$

At this point we will make the further assumption

$$S_{ij} = S_{ik}, \qquad i \neq j, \qquad i \neq k. \tag{14}$$

This assumption was not made in the previous section; however, it is reasonable for a resonant cavity TM_{010} combiner if adjacent lines are not close together. We advance the argument that at frequencies near resonance, the input coupling is primarily to the cavity mode, and any capacitive or inductive coupling directly between the input lines is, by comparison, negligible. Evidence exists for this assumption from measurements made on a four-way TM_{010} combining network [3].

Since $E_i^+ = -E_i^-$ on the shorted inputs, we can write

$$E_\Sigma^- = S_{i\Sigma}(E_1^+ - (N-2)E_i^-) \tag{15a}$$

$$E_j^- = S_{ij}(E_1^+ - (N-2)E_i^-) \tag{15b}$$

$$E_1^- = S_{11}E_1^+ - (N-2)S_{ji}E_i^- \tag{15c}$$

$$E_i^- = S_{ij}E_1^+ / (1 + S_{11} + (N-3)S_{ij}) \tag{16}$$

where the signal input is at port 1, and the Σ and jth ports are matched.

Substitution of (16) into (15) leads to

$$b_{11} = S_{11} - (N-2)S_{j1}^2 / (1 + S_{11} + (N-3)S_{j1}) \tag{17a}$$

$$b_{j1} = S_{j1} - (N-2)S_{j1}^2 / (1 + S_{11} + (N-3)S_{j1}) \tag{17b}$$

$$b_{\Sigma 1} = S_{\Sigma 1} - (N-2)S_{\Sigma 1}S_{j1} / (1 + S_{11} + (N-3)S_{j1}). \tag{17c}$$

Thus, there are three knowns: b_{11}, b_{j1}, and $b_{\Sigma 1}$ on the left, and the desired three unknown S-parameters on the right. With some ingenuity and algebraic manipulation, (17) can be inverted giving

$$S_{j1} = b_{j1}(1 + b_{11} - b_{j1}) / (1 - (N-1)b_{j1} + b_{11}) \tag{18a}$$

$$S_{\Sigma 1} = b_{\Sigma 1}(1 + b_{11} - b_{j1}) / (1 - (N-1)b_{j1} + b_{11}) \tag{18b}$$

$$S_{11} = b_{11} + (N-2)b_{j1}^2 / (1 - (N-1)b_{j1} + b_{11}). \tag{18c}$$

This leaves only $S_{\Sigma\Sigma}$, a quantity that is not usually required since the output port is normally matched, undetermined. For the sake of completeness, however, a signal input can be fed into the sum port, with ports 1 and j terminated. Following the same general derivation

$$b_{\Sigma\Sigma} = S_{\Sigma\Sigma} - (N-2)S_{\Sigma 1}^2 / (1 + S_{11} + (N-3)S_{j1}) \tag{19}$$

and inverting with the aid of (17)

$$S_{\Sigma\Sigma} = b_{\Sigma\Sigma} + (N-2)b_{\Sigma 1}^2 / (1 - (N-1)b_{j1} + b_{11}). \tag{20}$$

C. The Degradation Problem

Ideally, when a source fails or burns out in a combiner such as the one in Fig. 1, the output power would be reduced only by the amount contributed by that source. That is to say, if M sources fail in an N-way combiner, the resulting output power would be multiplied by the fraction

$$ID = (N-M)/N. \tag{21}$$

Reported results [1], [4] however, have shown that degradation approximately follows the relation

$$RD = (N-M)^2 / N^2. \tag{22}$$

Equation (22) can be readily derived by assuming perfect isolation between the input ports of the matched combining network. However, coupling does exist, and its presence, as this section shows, creates a situation in which the degradation may be somewhat less than that given by (22), or it may be much more, but the situation is never as good as shown by (21).

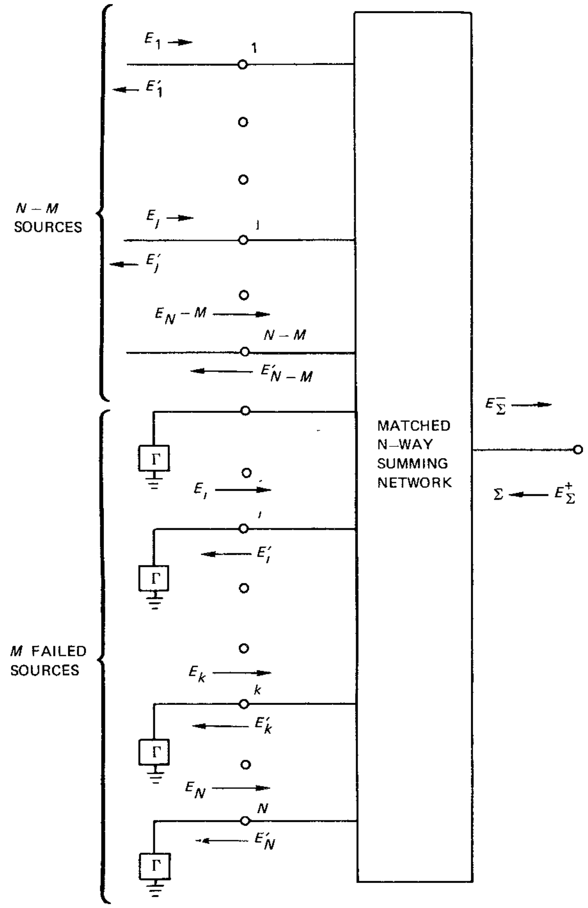


Fig. 6. N-way summing network block diagram model for graceful degradation.

Fig. 6 is a block diagram of the matched combiner—i.e., the matched combiner block includes everything inside of reference plane "B" in Fig. 2.

We assume that M of the sources have failed in an identical fashion so they have the same reflection coefficient Γ. It is also assumed that the remaining $N-M$ sources are independent of load conditions and continue to operate normally. If the sources are negative-resistance devices which generate power as a function of load impedance, this assumption is questionable. However, for a small percentage of failed devices (large N and small M), this effect should be negligible. Since the matched condition requires N identical inputs, we no longer have $E_j' = 0$.

There is a scattering matrix relation

$$(E') = (C)(E) \tag{23}$$

where the same conditions are assumed as previously. Since $E_k = \Gamma E_k'$, it follows that

$$E_\Sigma^- = C_{i\Sigma}\left[(N-M)E_j + M\Gamma E_k'\right] \tag{24}$$

where the subscript k refers to a "failed" input. Substitute

$$E_k' = \frac{(N-M)C_{ik}E_j}{1 - \Gamma\left[C_{kk} + (M-1)C_{ik}\right]}. \tag{25}$$

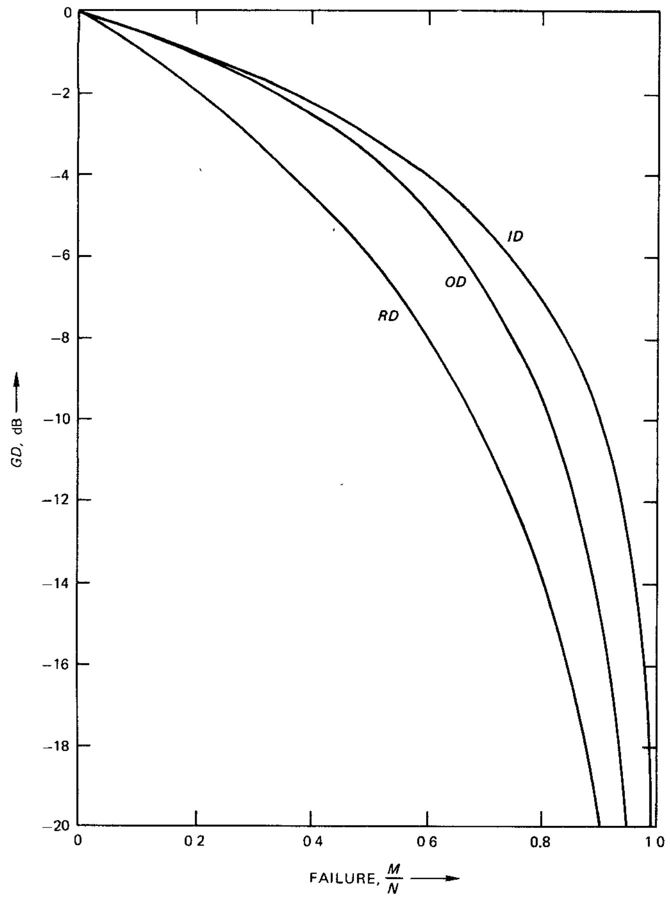


Fig. 7. Comparison between types of graceful degradation.

into (24), resulting in

$$E_{\Sigma}^{-}=C_{i\Sigma}(N-M)E_{j}\frac{1-\Gamma(C_{kk}-C_{ik})}{1-\Gamma[C_{kk}+(M-1)C_{ik}]}. \quad (26)$$

The ratio of the power output, after M diodes fail to the power out with no failures (the degradation factor), is given by

$$D=|E_{\Sigma M}^{-}|^{2}/|E_{\Sigma 0}^{-}|^{2}$$

$$D=\left(\frac{N-M}{N}\right)^{2}\frac{|1-\Gamma(C_{kk}-C_{ik})|^{2}}{|1-\Gamma[C_{kk}+(M-1)C_{ik}]|^{2}} \quad (27)$$

whereas, the actual efficiency is

$$\eta=(N-M)|C_{i\Sigma}|^{2}|1-\Gamma(C_{kk}-C_{ik})|^{2}/ |1-\Gamma[C_{kk}+(M-1)C_{ik}]|^{2}. \quad (28)$$

It has been assumed that when all of the sources are operating, they are matched; i.e., $E_j'=0$ in Fig. 3. It then follows

$$C_{jj}+(N-1)C_{ij}=0. \quad (29)$$

Substitution of (29) into (27) and (28) results in

$$D=\left(\frac{N-M}{N}\right)^{2}|1+\Gamma NC_{ik}|^{2}/|1+\Gamma(N-M)C_{ik}|^{2} \quad (30)$$

$$\eta=(N-M)|C_{i\Sigma}|^{2}|1+\Gamma NC_{ik}|^{2}/|1+\Gamma(N-M)C_{ik}|^{2}. \quad (31)$$

The failure of M sources leads to a nonzero reflection coefficient at the matching network of the remaining sources

$$R=-MC_{ik}/[1+\Gamma(N-M)C_{ik}]. \quad (32)$$

For a matched reciprocal N-way summing network, with efficiencies approaching 100 percent, it can be shown [3] $|C_{ik}|\cong 1/N$. Let ϕ be the phase angle of C_{ik}. (In this connection, it is interesting to note that while we do not know what this phase angle is, (29) requires that C_{ij} be 180° out of phase with C_{jj}.) If we assume that failure of a source introduces no additional losses (e.g., it fails as an open or a short), then $\Gamma=e^{j\theta}$. Substitution of these values into (30) results in

$$D=\left(\frac{N-M}{N}\right)^{2}|1+e^{j\gamma}|^{2}/\left|1+\left(\frac{N-M}{N}\right)e^{j\gamma}\right|^{2} \quad (33)$$

where $\gamma=\theta+\phi$. If $\gamma=2\pi$, degradation is minimized and is given by

$$OD=4(N-M)^{2}/(2N-M)^{2}. \quad (34)$$

This equation only represents an upper boundary of graceful degradation for solid-state power combiners and indicates that with careful circuit design, there may be room for improvement over previous results [1], [5]. Comparisons of (21), (22), and (34) are shown in Fig. 7. The results of (34) also agree with those published earlier by Saleh [6].

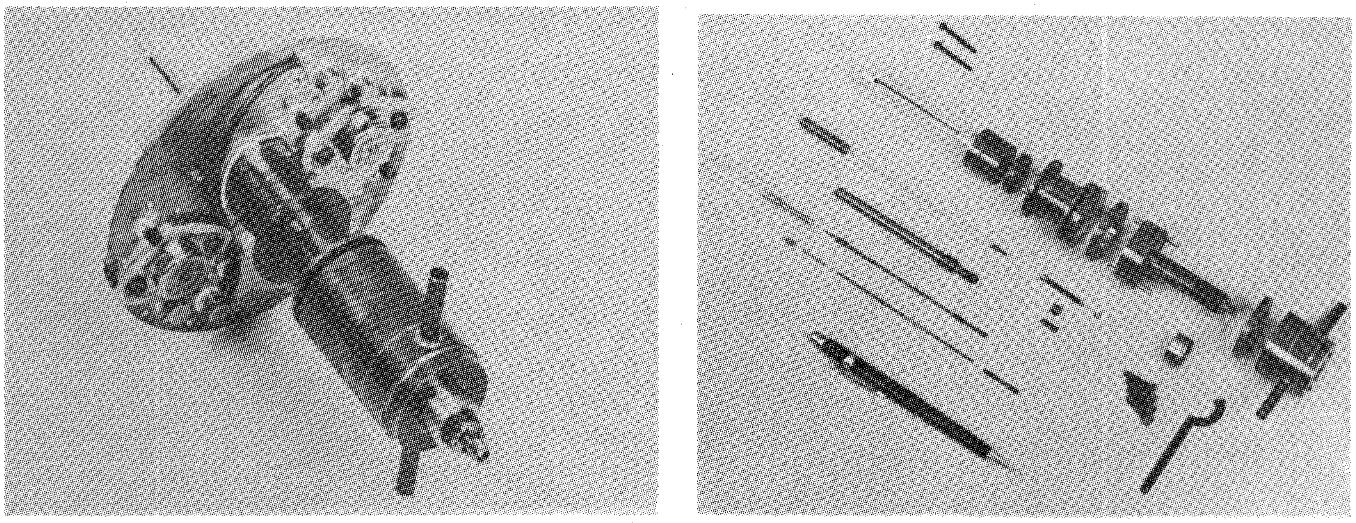
III. Measurements

In order to characterize a network in terms of efficiency and bandwidth, modified versions of (11) and (12) were programmed into an HP9825 desk calculator, which is part of a semiautomatic network analyzer (SANA). The two equations can be greatly simplified by assuming perfect symmetry ($S_{ij}=S_{ik}$ for all combinations of j and k from 1 to N except j or $k=i$), which is a good approximation for TM_{010} combining networks if the adjacent lines are not close together [3]. The modified equations then only require the measurement of three S-parameters to estimate efficiency and impedance

$$\eta\cong\frac{N|S_{\Sigma i}|^{2}}{1-|S_{ii}+(N-1)S_{ji}|^{2}} \quad (35)$$

$$\frac{Z_{ic}}{Z_{0}}\cong\frac{1+[S_{ii}+(N-1)S_{ji}]}{1-[S_{ii}+(N-1)S_{ji}]}. \quad (36)$$

Initially, a two-way TM_{010} network (Fig. 8) was optimized for efficiency on the SANA. The network has 4 degrees of tunability which consist of a variable-depth coupling probe and tuning slug, both of which are located at the center of the flat walls of the cavity and opposite each other. Initially, the cavity diameter was cut below the estimated value for the desired center frequency and gradually machined larger. Finally, tunable flat stabilizing loads were mounted at the end of each coaxial line and adjusted to the same position to maintain the required symmetry. After optimization, the network has the potential of com-



(a) (b)

Fig. 8. (a) Two-way TM_{010} network implemented as a power combiner. (b) Two-way power combiner disassembled.

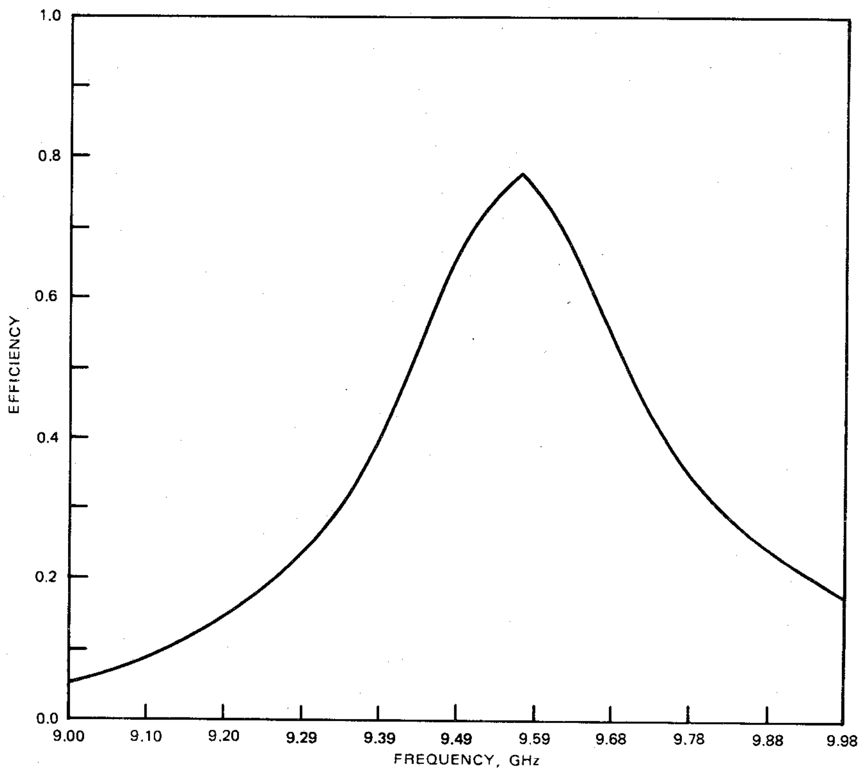


Fig. 9. Potential efficiency of a two-way combining network.

bining two devices at almost 80-percent efficiency (Fig. 9) if properly impedance matched. The impedance to be matched (Fig. 10) is referenced to the center of the cavity. It is important to note that any matching networks would have to be moved back along the coaxial lines in order not to modify the S-parameters by perturbing the cavity/coaxial line fringing fields.

Measurements were also made on a 16-way TM_{010} network (Fig. 11) which has the same tuning parameters as the two-way unit. To prepare this unit for testing, 14 of the input ports were terminated in internal tapered loads. Optimization produced a potential efficiency of almost 90

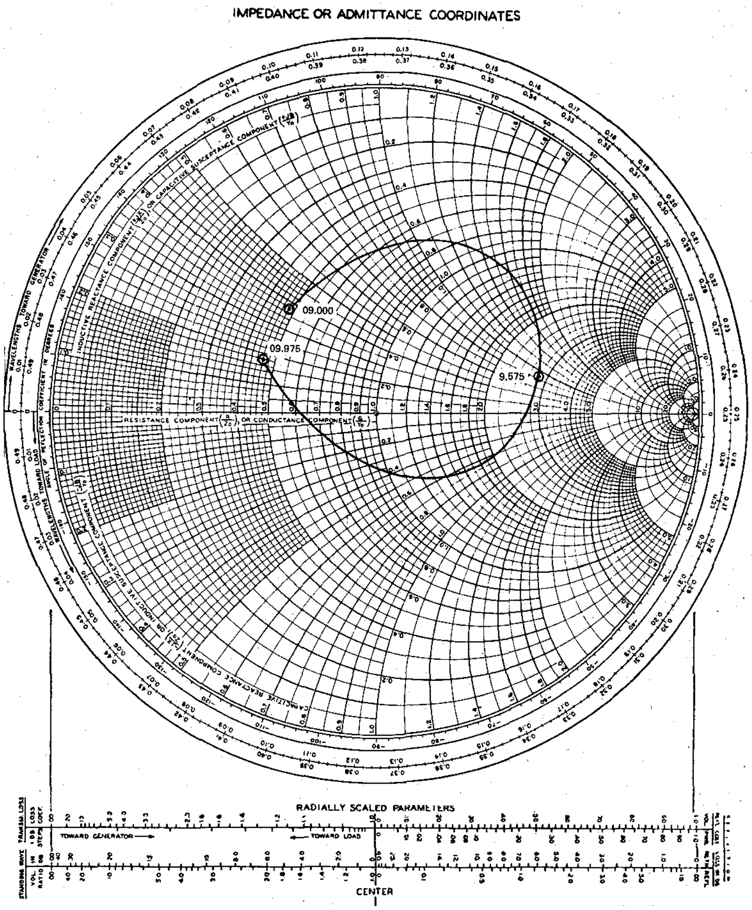
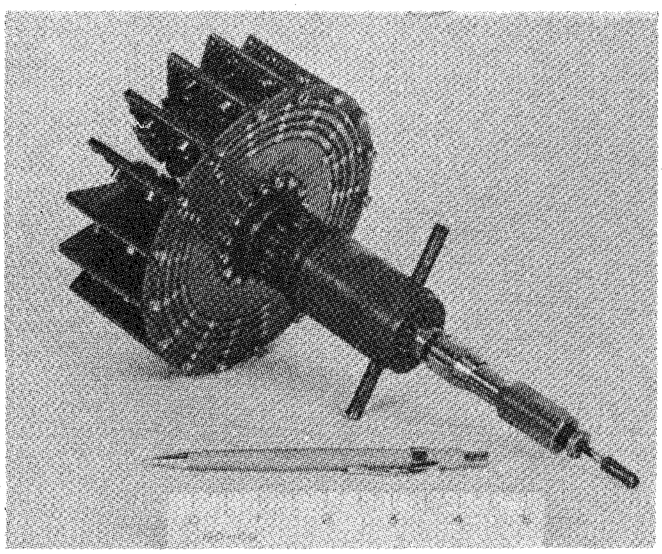
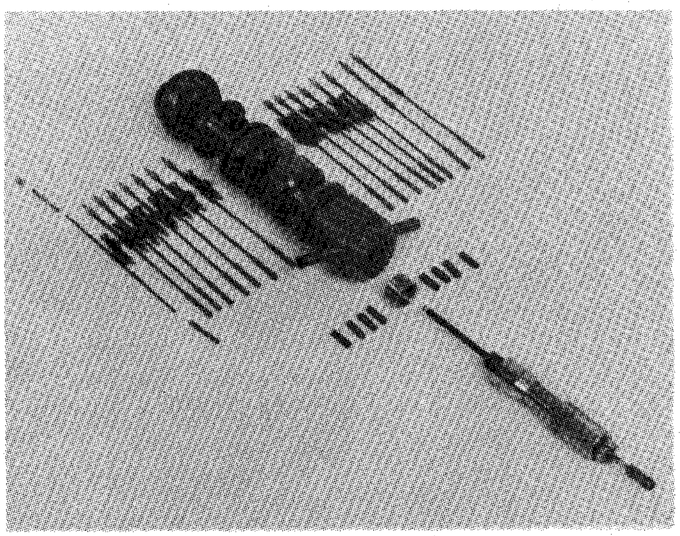


Fig. 10. Midcavity input impedance of a two-way combining network.



(a)



(b)

Fig. 11. (a) Sixteen-way TM_{010} network implemented as a power combiner. (b) Sixteen-way power combiner disassembled.

percent (Fig. 12) if properly matched (Fig. 13). The 16-way network showed better efficiency and broader bandwidth, as predicted by theory [2]. No comprehensive error analysis has yet been attempted that would determine efficiency and impedance accuracy; however, from observing (35) and (36), it can be seen that as the composite reflection

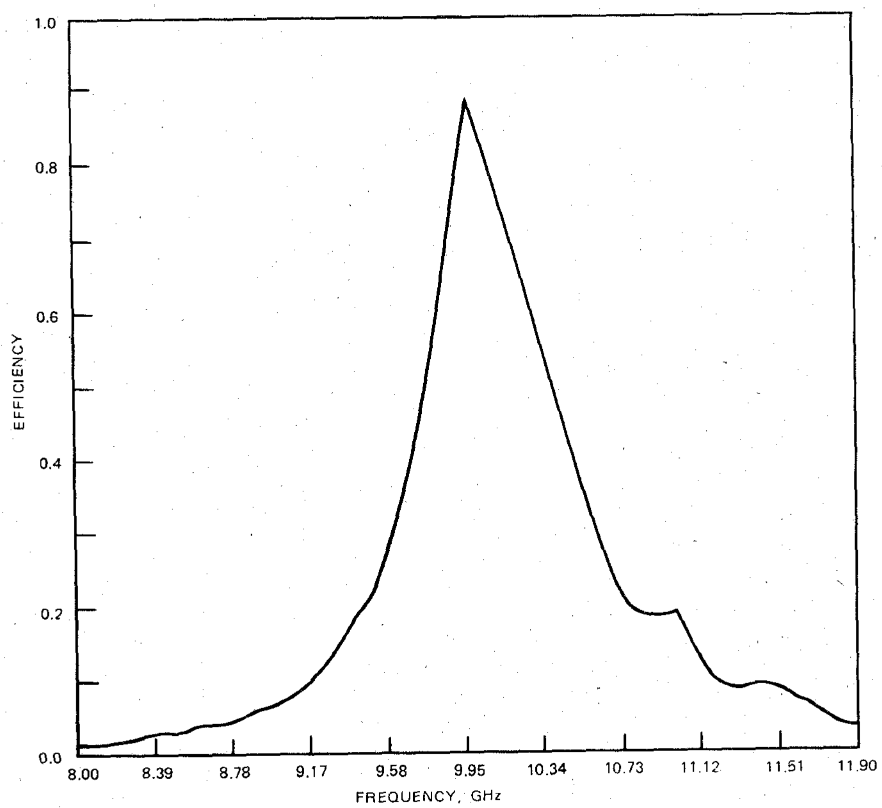


Fig. 12. Potential efficiency of a 16-way combining network.

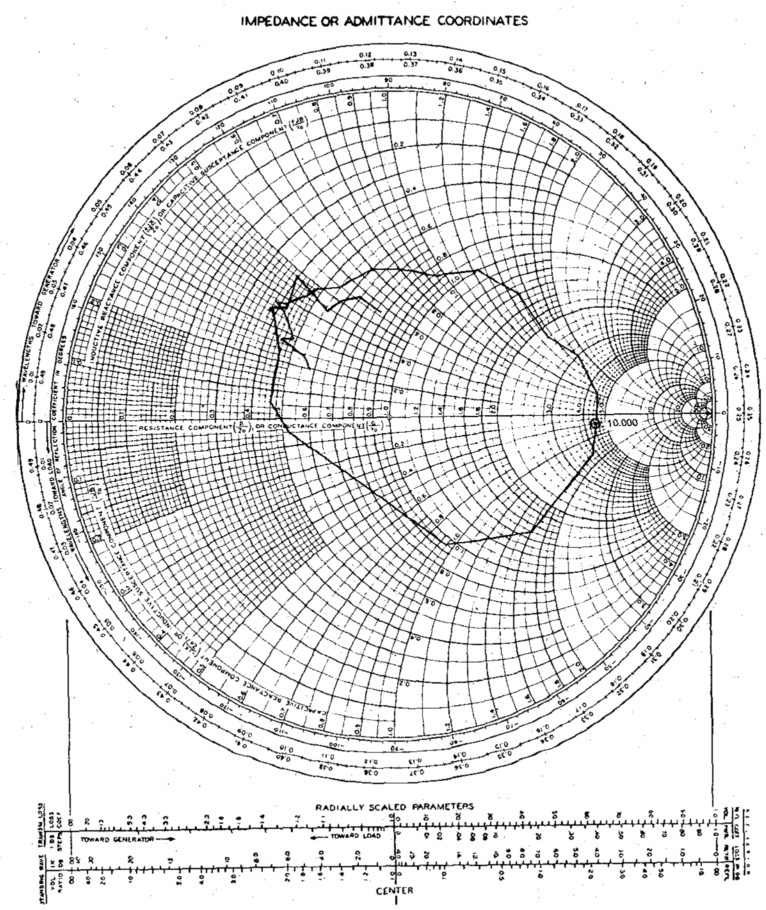


Fig. 13. Midcavity input impedance of a 16-way combining network.

coefficient $|S_L|$ approaches one, large errors are possible, especially in the efficiency equation. For small $|S_L|$, efficiency error is due mainly to the sum port transmission coefficient ($S_{\Sigma j}$), which the SANA can measure to an accuracy of ± 3 percent or less.

IV. Conclusions

Theory has been developed to calculate the potential efficiency and input impedance of symmetrical N-way summing networks. A modified version of these equations, which assumes perfect symmetry, was programmed into a network analyzer to create a new measuring instrument [7], that allows power-combiner circuitry to be optimized separately from the sources that are to be combined. In addition, theory was developed to predict the graceful degradation effects of perfectly symmetrical N-way power combiners, which indicates that there may be room for improvement in this area if the proper design techniques are used.

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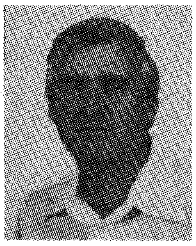
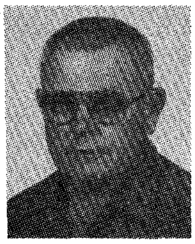


Darry M. Kinman (M'61) graduated from Fresno State College in 1960, receiving the B.S.E.E. degree. He was awarded the M.S.E.E. in 1968 from the University of California, Los Angeles.

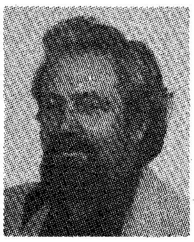
From June 1960 to the present time he has been employed by the Naval Weapons Center, China Lake, CA, where he has been involved in microwave circuit and subsystem design for guided missile data links and active radar seekers. Recently he has been involved in microwave power combining circuit development.

Mr. Kinman is a member of the Microwave Theory and Technique Professional Group.



David J. White (M'74) was born in Fallbrook, CA, in 1929. From 1949 to 1952 he was in the U.S. Army, attending radar repair school at Fort Monmouth, New Jersey and becoming radar instructor. He received the B.A. degree in physics in 1956 from the University of California, Riverside.

From 1956 to 1970 he was employed at the Naval Ordnance Laboratory, Corona, CA. From 1970 to the present he has been with the Naval Weapons Center, China Lake, CA, as a Research Physicist. During his career he has been involved in the microwave measurements of the properties of ferroelectrics, faraday rotation and magnetoresistance in semiconductors, pyroelectric detectors, surface acoustic wave devices, and mathematical analysis of microwave networks. Recently he has been engaged in work on artificial dielectrics and radome materials and design.



Marko Afendykiw (S'57–M'74) received the B.S. degree in electrical engineering from Wayne State University, Detroit, MI, in 1957, the M.S.E. degree and the Degree of Engineer, both in electrical engineering, from the University of Michigan, Ann Arbor, MI, in 1963 and 1971, respectively.

Since 1957 he has been with the Naval Weapons Center, China Lake, CA, where he worked on design and development of satellite tracking systems, passive radar, transreflector antennas, and passive and active microwave circuits. Currently he is directing the design and development of microwave solid-state sources for applications in missile seeker transmitters.

Mr. Afendykiw is a member of Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.