a desired effect. By selecting a greater radius of the ring, a larger volume would be heated.

IV. Conclusions

The design formulae and experimental results of a 2.45-GHz ring-type microstrip radiator for potential hyperthermia applications are given. Microstrip radiators offer the advantage of being small, lightweight and capable of conforming to the shape of the body. Cooling of the skin is easily accomplished by the air flow through a gap between the skin and the radiator. The preliminary results obtained for a radiator operating at 2.45 GHz are encouraging. A defined volume of muscle can be heated without overheating skin and fat. The energy can be effectively coupled to the tissue, with a relatively small leakage (1 mW/cm^2 per 10 W of net power delivered to the tissue). The heated volume can be controlled by the radiator dimensions. The radiator was designed and tested at 2.45 GHz, because of availability of the test instrumentation, however dimensionally scaled radiators at other frequencies can be designed using the principles described. If at the same time a small cross section of the heated volume is required, a radiator having a substrate of higher dielectric constant can be used.

References

[1] M. F. Iskander and C. H. Durney, "Electromagnetic techniques for medical diagnostics: A Review," *Proc. IEEE*, vol. 68, pp. 126–132, Jan. 1980.

[2] F. Sterzer *et al*, "Microwave apparatus for the treatment of cancer," *Microwave J.*, vol. 23, pp. 39–44, Jan. 1980.

[3] M. F. Iskander and C. H. Durney, "An electromagnetic energy coupler for medical applications," *Proc. IEEE*, vol. 67, pp. 1463–1465, Oct. 1979.

[4] I. Wolff and N. Knoppik, "Microstrip ring resonator and dispersion measurements on microstrip lines," *Electron Lett.*, vol. 7, pp. 779–781, Dec. 1971.

[5] Y. S. Wu and F. J. Rosenbaum, "Mode chart for microstrip ring resonators," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-21, pp. 487–489, July 1973.

[6] I. J. Bahl and S. S. Stuchly, "Analysis of a microstrip covered with a lossy dielectric," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-28, pp. 104–109, Feb. 1980.

[7] E. Yamashita and R. Mittra, "Variational method for the analysis of microstrip lines," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-16, pp. 251–256, Apr. 1968.

[8] A. W. Guy, "Analyses of electromagnetic fields induced in biological tissues by thermographic studies on equivalent phantom models," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-19, pp. 205–215, Feb. 1971.

The Traveling-Wave Divider/Combiner

ALAIN G. BERT AND DIDIER KAMINSKY

***Abstract*—A new kind of distributed power divider/combiner circuit for use in octave (or more) bandwidth microstrip power transistor amplifiers is presented.**

The design, characteristics, and advantages are discussed. Experimental results on a four-way divider and divider/combiner are presented and compared with theory.

First experimental results on a six-way traveling-wave divider (TWD) are also presented.

Introduction

Several watts per chip are now obtainable at X-band with power FET's, due to recent improvements in their technology, yet bandwidths remain relatively narrow (20 percent typical). Wider bandwidths may be achieved by the use of internal matching techniques [1]. Nevertheless, the greater the power, the smaller is the input impedance and bandwidth. Therefore, large power values (10 W or more) over octave bandwidths for ECM applications, require wide-band combining circuits.

It must, however, be kept in mind when speaking of combiners, that high-power FET's are already the result of internal power combining. At the chip level, individual FET's are connected in parallel to form a power module. Then a certain number of such modules (generally 2 or 4) are bonded to some capacitances which are part of the input or output matching circuits [2].

Apart from the fact that the input impedance decreases with the number of elements connected in parallel, another reason that there is a limitation in the power of each power FET is related to its longitudinal dimension which becomes comparable to a quarter of a propagation wavelength on the GaAs substrate, particularly when the search for more power is associated with the trend toward high frequencies. A typical value for this longitudinal dimension is 500 μm/W of RF power while a quarter of a

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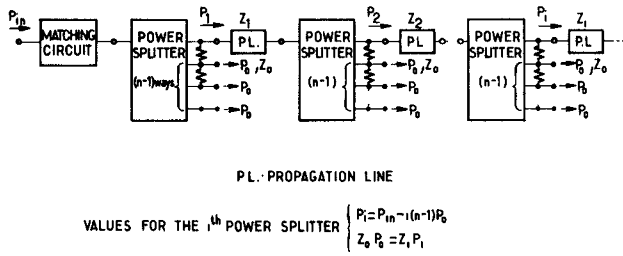


Fig. 1. Basic block diagram for a traveling-wave divider (TWD).

wavelength for a 50-Ω microstrip line on GaAs is 1600 μm at 16 GHz. Therefore the input power distribution between the individual FET's can hardly be uniform for multiwatt power levels when the frequency is increased to the *Ku*-band, unless the propagation effect is taken into account in the power distribution structure.

A circuit will be described that not only achieves this goal, but at the same time offers the possibility of increased bandwidth.

Let us start with N medium power modules, which are connected by combining circuits to an input generator and an output load. In such a case, it can be readily appreciated that phase lags between the input ports of successive modules are of no consequence on power addition provided:

1) the input powers are equal,
2) the phase differences are compensated in the combining circuit.

However for $N>2$ most of the amplifier combiners are realized with in-phase dividers [3]–[5], even in recent monolithic multi-FET circuits [6]. Such a choice prevents the designer from taking advantage of the degree of freedom that is given by the multiplicity of the amplifying paths. A particularly difficult problem is that in-phase dividers, even with bridging resistors, do not absorb the waves simultaneously reflected by identical mismatched loads. Frequency-selective mismatches at the input of the FET's are useful in wide-bandwidth operation, due to the 6-dB/octave gain slope decrease and the reflected power at the lower frequencies must be absorbed somewhere.

If the reflected wave is not absorbed in the amplifier, some solution must be found to absorb it into a dummy load outside the amplifier [4]. If not, resistive matching, quadrupoles must be introduced at the inputs of each FET [8], which makes the circuit more complicated and reduces the gain–bandwidth product. The traveling-wave divider (TWD) that will be described, allows the waves reflected from identical mismatches to be absorbed within the dividing structure.

Traveling-Wave Power Dividers

The basic block diagram of a TWD is shown on Fig. 1. It results from cascading in-phase n-way power splitters connected by sections of propagation lines in such a way that incident waves have different times of arrival at the output ports of successive power splitters. More complicated structures can be derived by connecting blocks of the same type to the outputs of a first divider, and so on. Impedance values at the outputs of each power splitter are chosen proportional to the reciprocal of the circulating powers. Therefore, in the absence of reflected waves, the voltages are equal at the outputs of each power splitter and no power is absorbed through the resistors connected between these outputs. The forward traveling-wave propagation is therefore unaffected. If the TWD is connected to identical output loads which are no longer matched, reflected waves are generated that are out of phase at the output ports of the power splitters provided the phase angle staggering of the propagation line is different from 0 or π. Currents circulate through the resistors so that reflected power can be absorbed.

Bridging resistors are necessarily introduced between the outputs of each power splitter but other resistors may be introduced between any other points of the circuit provided their RF potentials are equal for matched conditions. This may help in increasing the isolation between output ports of the TWD over a wide bandwidth.

Traveling-Wave Power Combiners

An n-way TWD can be used as a combiner if n power sources are connected to the n ways of the TWD. Power addition will be obtained at the main port provided the phase lags between the sources are the opposite of what they would be between output ports when used as a divider. Such a situation can be clearly obtained if the same TWD is used as a divider which feeds the combiner through identical paths.

Four-Way MIC Traveling-Wave Dividers (TWD)

Numerous trees can be generated according to the number of power splitters and the number of outputs of each power splitter. We will restrict ourselves to TWD's having only binary power splitting ($n=2$). The trees may be then classified in two classes:

Class I: each output of a power splitter feeds another power splitter;
Class II: only one output feeds another power splitter, the other output feeding one load.

In designing a TWD, a mixture of the two classes may be used. The number of trees increases rapidly with the number of outputs: 2 for a four-way, 6 for a six-way, 23 for an eight-way, $\cdots$.

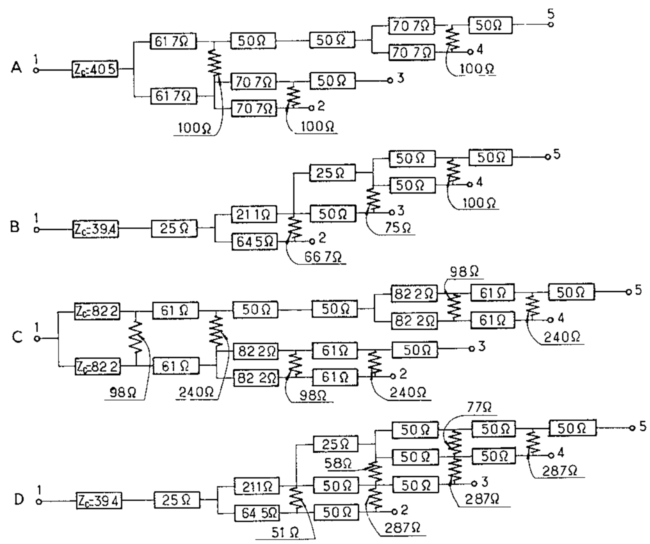


Fig. 2. Four-way TWD.

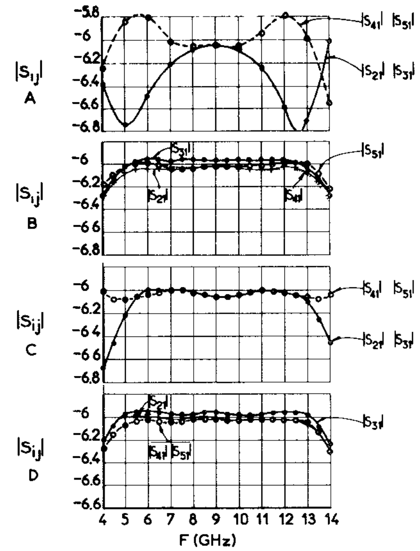


Fig. 3. Transmission parameters of a four-way TWD (dB).

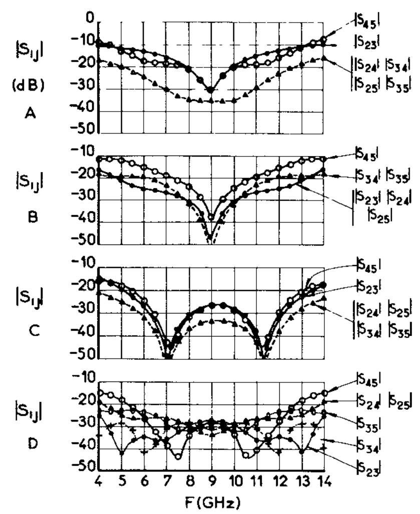


Fig. 4. Isolation between output ports of a four-way TWD.

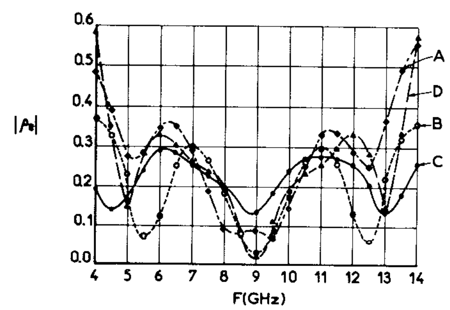


Fig. 5. Input reflection coefficient of a four-way TWD under short-circuit conditions.

On Fig. 2, circuits A and B illustrate the two classes for a four-way TWD in their simplest version. Circuits C and D are of the same type as A and B, respectively, except for the number of bridging resistors to increase the isolation between output ports. Each box represents a microstrip line.

In structure A and C, all the power splitters are symmetrical two-way Wilkinson dividers which include the impedance transformation [9]. In structure A, to have a better VSWR, the first divider is a two-stage power splitter. The impedance values of the microstrip lines for these two circuits are higher than the load and source impedances and all the microstrip lines are a quarter wave in length.

Structures B and D correspond to the most unsymmetrical case (Class II). The resistors are connected between adjacent lines. The topology of the structures must take into account the differences between the velocities due to the dependence of the ϵ_{eff} with the impedance values. All circuits have colinear equally spaced outputs.

In structures B and D, the impedance transformations are performed at the input port of the structure. The impedance values of the microstrip lines in these circuits are mostly lower than the load and source impedances.

In structures B and D presented here, the first power splitter is included in the three-step quarter-wave impedance transformer to increase the impedance values. For such a four-way octave bandwidth TWD all impedances values remain between 21 and 65 Ω.

In structures A and C all lines are quarter wave in length since they perform impedance transformations. In structures B and D, the lines may be chosen different from a quarter-wave length, except for the input matching circuit, giving the possibility of combining in a great number of ways within a small surface area.

A more precise comparison can be made from the computed scattering parameters of structures A, B, C, and D as they appear in Figs. 3–5. Here, all the microstrip lines are a quarter wave in length around 9 GHz. The computations have been made using a CAD program which performs multipole analysis (giving the multipole S-matrix) and quadrupole optimization. One particular feature of this program is its ability to optimize circuit parameters for equalizing powers delivered at different points of the circuit. This optimization has not been used here because of the straight-forward synthesis procedure for circuits A, B, C, and D, but is of great interest for more intricate structures. Coupling between adjacent microstrip lines has not been taken into account here, for the sake of simplicity. Only variations of ϵ_{eff} with frequency were introduced.

It is important to underline the difference in transmission behavior between TWD's of types A or C and TWD's of types B or D. The power values are much more equal out of the four-ways in the last types of circuit for very wide bandwidths. The isolation between output ports is always limited by the characteristics of the last power splitter. Circuit B gives better results than circuit A. Circuits C and D give similar results. Input VSWR for the matched condition is greater for Wilkinson-type circuits (1.25 for circuit C instead of 1.13 for circuits B and D), their output reflection coefficients, however, are much lower. Fig. 5 shows the input reflection coefficient when all outputs are shorted which shows a minimum return loss values of about 10 dB over more than an octave bandwidth; rather surprisingly, the general behavior of the curves is very similar for circuits A, B, C, D although they have different numbers of resistors. These results lead to the following conclusions.

1) As far as transmission is concerned, circuit B is a good choice (wide band, simplicity, good equalization).

2) All circuits exhibit almost equivalent return loss under short-circuit conditions.

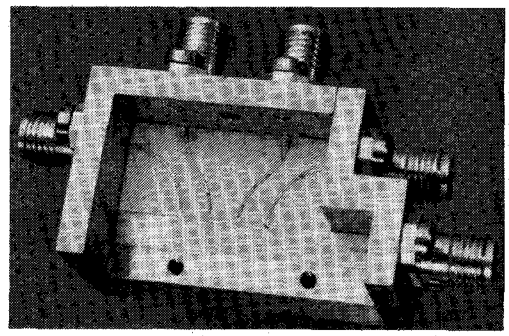
3) If a minimum value of 15 dB over octave bandwidth is sufficient for the isolation between any output port, circuit B should be preferred.

4) If the reflection coefficients must be small from all the output ports over an octave bandwidth, then circuits A, C, or D are better.

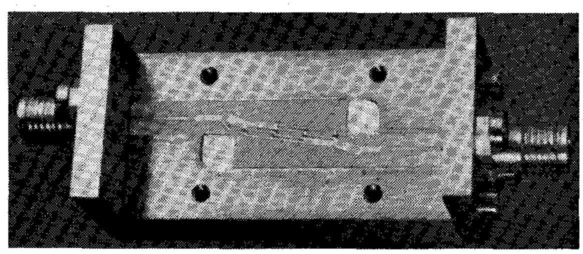
Experimental Performance of a Four-Way Divider

Fig. 6(a) is a photograph of a four-way TWD of type B. The input impedance transformer was calculated to cover the 6–12-GHz band [10]. The distance between successive output ports is 3.4 mm. The circuit is fabricated on a 0.25-mm-thick alumina substrate. A thin substrate leads to smaller widths for the microstrip lines used and this has distinct advantages at high frequencies. Three thin-film resistors are present between adjacent strips to provide absorption of the reflected power. Each section of microstrip line is a quarter of a wavelength at a center frequency of 9 GHz. The first two-way power splitter is incorporated in the three-step input impedance transformer. The theoretical values of the quarter-wave impedance sections are given in Fig. 2. They correspond to microstrip lines of different widths and lengths, but special care must be taken to design the junctions between these different lines to obtain the desired power division.

Experimental results are shown on Figs. 7–9. The TWD feeds 4×50-Ω microstrips of equal lengths and measurements were made through SMA connectors. Transmitted powers in the four-ways were equal to within ±0.6 dB. The absorption of the reflected waves by the resistors of



(a)



(b)

Fig. 6. (a) Experimental design of a four-way TWD (0.25-mm-thick alumina). (b) Experimental dividing/combining structure (0.25-mm thick alumina).

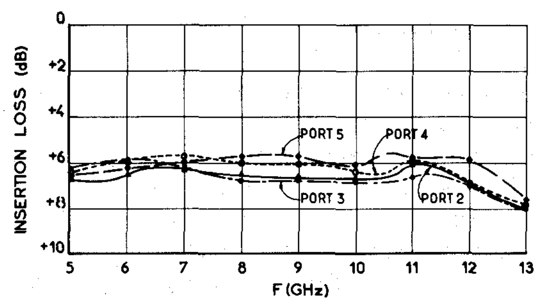


Fig. 7. Insertion loss between input port and output ports (measured).

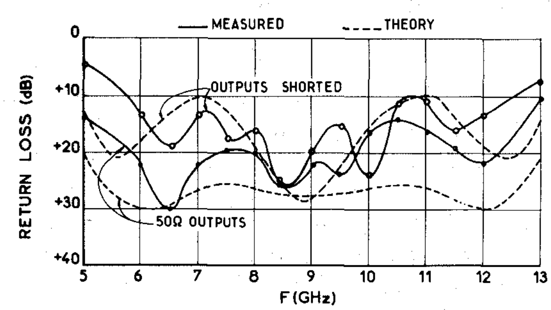


Fig. 8. Input return loss.

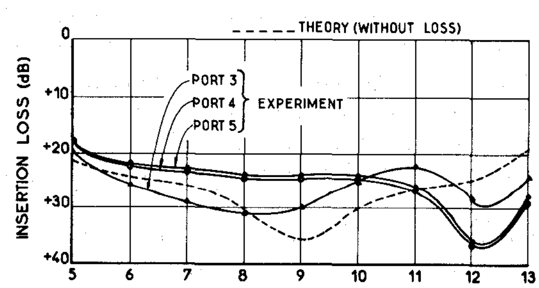


Fig. 9. Isolation between port 2 and other output ports.

the divider can be appreciated from Fig. 7. When all output ports are shorted, the input return loss remains greater than 10 dB. The isolation between port 2 and the other output ports is better than 21 dB over more than an octave bandwidth (6–12 GHz) and better than 13 dB for the three remaining ports.

Fig. 6(a) shows the advantage of a TWD circuit over in-phase Wilkinson-type circuits as far as the overall dimensions are concerned. In general, for the divider/combiner amplifiers, longitudinal dimensions (i.e., distance between successive ports) are imposed by the lateral dimension of each individual amplifier, for which a reasonable value corresponds to a quarter of a wavelength for most common cases. This four-way TWD spans no more than 2.5 mm transveral width.

The discrepancy with the theory (Figs. 8 and 9) in the upper part of the bandwidth is mainly due to the transition between the second and third sections of the input impedance transformer and to the sharp bends at the output ports.

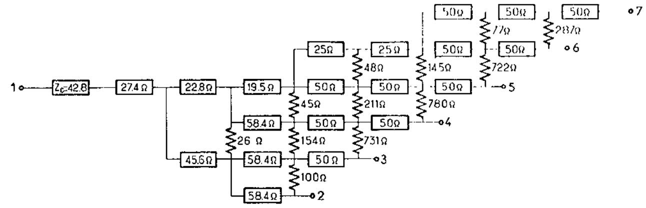
A better fit has been obtained with a circuit which has been fabricated on a 0.635-mm-thick alumina, with a larger transversal width. The model used gives a good description of what has been observed although the coupling between adjacent microstrip lines has not been taken into account.

A dividing/combining structure has also been measured, each output port of the first four-way TWD being connected to the corresponding "input" port of the second TWD working as a combiner (Fig. 6(b)). The input VSWR was better than 1.3 within the octave bandwidth and the total transmission power loss was 1 dB at 12 GHz and 0.5 dB at 6 GHz. The computer model of this dividing/combining structure shows that a phase deviation of 40° in any one of the four ways at 12 GHz results in an increased insertion loss of 0.4 dB.

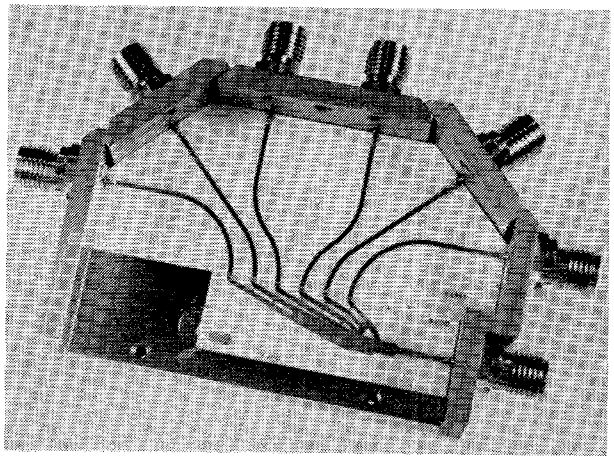
Six-Way Traveling-Wave Divider (TWD)

The basic block diagram of a six-way TWD is presented in Fig. 10(a) and the photograph of this first design in Fig. 10(b). The first power splitters have been included in the impedance transformer and the two first output ports have been individually separated from the main line to keep the characteristics impedances of all the microstrip lines to between 20 and 65 Ω. The thickness of the alumina is 0.635 mm.

This structure has also been computed on the same program to get the total S-matrix parameters. However, it is clear that the modeling of such a structure by simply connecting microstrip lines of different lengths and impedances can only give a rough description of reality. There is a need for a computer model where one could introduce the shape of such an intricate metallic design, and then given the excitation conditions, would end up with the RF electrical values and the Poynting vector anywhere under the circuit. This will become necessary for modeling monolithic circuits. The experimental results seem in quite reasonable agreement given the limitations



(a)



(b)

Fig. 10. (a) Block diagram of a six-way TWD. (b) First design of a six-way TWD.

just described. The insertion loss between the input port and the six output ports is 8 dB ± 1.4 dB from 2 to 12 GHz. The power output increasing gradually from ports 2 to 7, the compensation of this effect in a new design should not to be too difficult. The input VSWR under matched condition is less than 1.2 from 6 to 12 GHz, and is equal to 2.5 at 2 GHz. Between ports 2 and 3, the isolation is better than 10 dB from 2 to 14 GHz and better than 13 dB above 6 GHz. Between any other combination of ports, the isolation is always better than 16 dB from 3 to 14 GHz. Under short-circuit conditions, the input VSWR is less than 1.8 from 6 to 14 GHz, being equal to 3 at 2 GHz.

Conclusion

Although the TWD's which have been described have not yet been optimized, experimental results confirm the usefulness of this new type of power divider for high-power wide-bandwidth integrated FET power amplifiers.

The CAD program that has been used for designing these structures provides S-matrix coefficients which are in good agreement with the experimental results.

The wide-bandwidth capability can best be illustrated by giving some figures from the modeling of a 6–18 GHz, four-way TWD of type D circuit.

Over the whole 6–18-GHz bandwidth the input VSWR is less than 1.18 under matched conditions, isolation between any of the output ports is better than 18 dB, and the reflection coefficients are lower than 0.15 seen from the outputs. For a greater number of outputs, several individual TWD's may be fed by a first divider of the same type so as to maintain realizable strip-line impedance values, and all the output ports can remain in line.

Transistor amplifiers may be connected to the outputs of the TWD. They only need to be adjusted to have equal gains over the bandwidth for whatever input VSWR: reflected waves will be absorbed in the TWD. The greater the number of ways, the smaller will be the resulting input-reflection coefficient. Combining the powers from individual amplifiers will be effected by a similar TWD connected to the outputs of each amplifier: the total electrical path will then be the same whatever the path (as is the dividing/combining structure of Fig. 6(b)). This results in a compact all-planar MIC power amplifier.

Another interesting feature may be expected from such a traveling-wave amplifier when the output load turns out to be highly mismatched. Power, in this case, will be reflected back from the output and divided before reaching the outputs of the different amplifiers. Clearly, the phase of the reflection coefficients seen from the successive amplifiers will be different, therefore the amplitude and phase of their output signals will be modified compared to the previously matched condition. Currents will then circulate through the bridging resistors and this will limit the effect of the mismatched load.

Due to its compact topology and to its matching properties, the traveling-wave amplifier concept may be useful for a monolithic design. A GaAs semi-insulating substrate of 100 μm would lead to much smaller transversal dimensions.

Furthermore, a $\pi/2$ phase lag between the power splitters is not necessary for these TWD's, especially of types B or D, for the input reflected waves to be absorbed within the resistors. This effect will be essentially governed by the total longitudinal phase difference for whatever number of amplifiers used and so can be easily optimized.

Acknowledgment

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References

[1] K. Honjo, Y. Takayama, and A. Higashisaka, "Broad-band internal matching of microwave power GaAs MESFET's," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, no. 1, Jan. 1979.
[2] J. Srue, and Y. Takayama, "Ku and K-band internally matched high power GaAs FET amplifiers," *Electron. Lett.*, Aug. 30, 1979.
[3] N. Nagai, K. Ono, and E. Maekawa,"On some five-way hybrid power dividers," in *1978 Int. Microwave Symp. dig.*, p. 326.
[4] K. B. Niclas, "A 400 mW 11-18 GHz GaAs MESFET amplifier using planar power combining techniques," in *Proc. ISSCC*, p. 122, 1979.
[5] M. Cohn, B. D. Geller, and J. M. Schellenberg, "A 10 watt broadband FET combiner/amplifier," in *IEEE MTT Int. Microwave symp. Proc.*, p. 292, 1979.
[6] R. A. Pucel, "Power combiner performance of GaAs MESFET," *Microwave J.*, p. 51, Mar. 1980.
[7] K. Kurokawa, "Design theory of balanced transistor amplifiers," *Bell Syst. Tech. J.*, Oct. 1965.
[8] D. P. Hornbuckle and L. J. Kuhlman, "Broad-band medium power amplification in the 2-12.4 GHz range with GaAs MESFET's," *IEEE Trans Microwave Theory Tech.*, vol. MTT-24, no. 6, June 1976.
[9] S. B. Cohn, "A class of broadband three-port tem-mode hybrids," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-16, no. 2, Feb. 1968.
[10] L. Young, "Tables for cascaded homogeneous quarter wave transformers," *IRE Trans. Microwave Theory Tech.*, vol. MTT-7, p. 233–237, Apr. 1959.

A Least Squares Solution for Use in the Six-Port Measurement Technique

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Abstract—**Although based on the use of simple amplitude detectors, it is possible to obtain complex values of reflection coefficient, via the six-port technique, from the intersection of three circles in the complex plane. In a typical case, the circle centers are determined primarily by the six-port design and are nominally constant, while the radii are proportional to the square root of the ratio of the output of three of the detectors to a fourth one. As a practical matter, however, these circles will not intersect in a point because of noise or other errors in the detectors.**

This paper develops a procedure for choosing Γ in this context. Moreover, the question of what may be inferred about the system performance from the extent of this intersection failure is briefly considered.

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I. Introduction

THE SIX-PORT reflectometer, which uses four simple amplitude detectors, provides an alternative to the four-port reflectometer and complex ratio detector in im-