Intrinsic Attenuation*

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Summary—The name "intrinsic attenuation" is suggested for the concept originally called "intrinsic insertion loss" and reasons for the proposed change are given. Formulas are given to permit calculation of this quantity, given the scattering coefficients of the network. The use of lossless tuners to obtain a bilaterally matched nonreflecting network is described, and the expected reduction in attenuation is discussed. A graph is presented for rapidly estimating this reduction in attenuation, given minimum information about the network. Experimental results for badly mismatched coaxial attenuators show that the losses in the tuners may in some cases prevent one from realizing any reduction in attenuation.

Introduction

A CONCEPT called the "intrinsic insertion loss of a mismatched microwave network" was introduced[1] by Tomiyasu in 1955. Although this concept was defined and a method of measurement given, no explicit expression has been derived[2] for its calculation, given the parameters of the network. In addition the concept was developed to apply only to networks inserted in a matched (nonreflecting) system, and for which reciprocity applies. It has been stated[3] that the intrinsic insertion loss of a network is the one-way dissipative loss of that network, when actually it is the one-way dissipative loss of a combined network, including lossless matching transformers, adjusted to make the combined network nonreflecting.

It is the purpose of this paper to 1) propose renaming this concept "intrinsic attenuation" for reasons to be explained, 2) present formulas for its calculation in terms of the scattering coefficients of the network, 3) present a graph for quickly estimating the reduction in attenuation to be expected when tuning for bilateral match, given minimum information about the network, and 4) present some experimental data on the observed change in loss when tuning for bilateral matched (nonreflecting) conditions.

* Received December 3, 1962.

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[1] K. Tomiyasu, "Intrinsic insertion loss of a mismatched microwave network," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-3, pp. 40–44; January, 1955.

[2] An expression for the "essential or intrinsic loss of a network" is given as $L_I = -10 \log |S_{12}|^2$, eq. (11.12) on p. 465 of E. L. Ginzton, "Microwave Measurements," McGraw-Hill Book Company, Inc., New York, N. Y.; 1957. However the S_{12} in this expression is not a parameter of the network alone, but of a combined network including lossless matching transformers adjusted for bilateral nonreflecting matched conditions. One is not told how to obtain S_{21} from the parameters of the individual network.

[3] S. A. Rinkel and W. E. Waller, "Microwave Attenuation Measurements," PRD Rept., vol. 4, p. 6; April, 1955. Also see Ginzton.[2] Apparently no distinction is made between the one-way dissipative loss of the combined network and of the original network.

Expressions for Intrinsic Attenuation

An expression for the intrinsic attenuation of a microwave network in terms of its scattering coefficients may be derived from the results of the measurement method[1] of Tomiyasu. He has expressed the loss in terms of the radius R of a circular locus of reflection coefficient Γ_1 and the distance ρ of the center of this circle from the origin. One can express R and ρ in terms of the scattering coefficients of the microwave network, and obtain the following expression for the intrinsic attenuation A_I:

$$A_I = 8.686 \tanh^{-1} x = 10 \log_{10} \frac{1+x}{1-x},$$

where

$$x = \sqrt{\frac{(1-R)^2 - \rho^2}{(1+R)^2 - \rho^2}}, \tag{1}$$

and

$$R = \frac{|S_{21}|^2}{1 - |S_{22}|^2},$$

and

$$\rho = \left| S_{11} + \frac{S_{21}{}^2 S_{22}{}^*}{1 - |S_{22}|^2} \right|,$$

and the asterisk denotes that the complex conjugate is taken.

Although this expression is useful, it is restricted to microwave networks for which reciprocity holds in the form $S_{12} = S_{21}$. An alternate expression which is not so restricted and which relates the intrinsic attenuation to the maximum efficiency of the network can be obtained as follows.

Consider the circuit shown in Fig. 1 in which lossless tuners are placed on each side of a network connected between a generator and a load. Suppose that we are required to adjust the tuners so that maximum power is absorbed by the load.

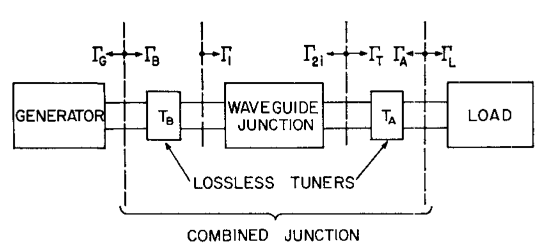

Fig. 1—Arrangement for tuning for minimum loss.

This could probably be done by a succession of alternate adjustments of the tuners, but a different method to be described is better for our purposes. First adjust[4] T_A so that the waveguide junction (microwave network) has maximum efficiency η_m. The efficiency η is defined to be the ratio of the net power output to the net power input. It is noted that since the efficiency is independent of the generator characteristics, this condition does not necessarily correspond to a conjugate match.

Then adjust T_B so that the generator delivers all of its available power to the input of the waveguide junction. This occurs when a conjugate match is obtained, and if the waveguide has a real characteristic impedance, as is the case for lossless waveguides, then $\Gamma_B = \Gamma_G^*$. It is intuitively evident that no further adjustments of the tuners can increase the power absorbed by the load and hence $\Gamma_{2i} = \Gamma_T^*$. Therefore the loss of the combined waveguide junction, including the tuners, is minimum.

A number of loss concepts could be applied here, but one will be chosen which leads to a simple mathematical expression for minimum loss. Consider the transducer loss L_T, which is defined to be the ratio expressed in decibels of the available generator power to the power absorbed by the load. It is a measure of how well, or how poorly, the waveguide junction performs as a transducer.

In the case of the properly adjusted combined network of Fig. 1, the power input equals the available power of the generator and therefore the ratio in question is the reciprocal of the efficiency of the combined junction. Since the tuners are lossless, this is also the efficiency of the waveguide junction under consideration. We recall that this efficiency had been maximized by the first adjustment and therefore the minimum transducer loss is

$$L_{TM} = 10 \log_{10} \frac{1}{\eta_m} \cdot \tag{2}$$

It can be shown[5,6] that

$$\eta_m = \frac{Z_{01}}{Z_{02}} \cdot \frac{|S_{21}|^2(1 - |\Gamma_{TM}|^2)}{|1 - S_{22}\Gamma_{TM}|^2 - |(S_{12}S_{21} - S_{11}S_{22})\Gamma_{TM} + S_{11}|^2}, \tag{3}$$

where S_{11}, S_{12}, S_{21}, and S_{22} are the scattering coefficients of the waveguide junction, Z_{01} and Z_{02} are the real characteristic impedances of arms 1 and 2, and

$$\Gamma_{TM} = \frac{B}{2A}\left(1 \pm \sqrt{1 - \frac{2|A|}{B}}\right), \tag{4}$$

where

$$A = S_{22} + S_{11}^*(S_{12}S_{21} - S_{11}S_{22}),$$

and

$$B = 1 - |S_{11}|^2 + |S_{22}|^2 - |S_{12}S_{21} - S_{11}S_{22}|^2.$$

The expression for Γ_{TM} can be obtained by solving the equation grad $\eta = 0$, or by another method.[7]

The expression obtained for the minimum transducer loss above contains no terms which depend upon generator or load characteristics, and hence is intrinsic to the waveguide junction. Since one can define attenuation as the transducer loss when $\Gamma_G = \Gamma_L = 0$, the minimum or intrinsic transducer loss equals the minimum or intrinsic attenuation. The above definition of attenuation is consistent with those in common usage. The use of insertion loss is avoided here because, in its most general interpretation, it depends upon the generator and load characteristics, and cannot therefore be regarded as intrinsic to a waveguide junction.

The Bilateral Nonreflecting Matched Condition

When adjusting tuners T_A and T_B shown in Fig. 1 for minimum loss, one does not necessarily obtain a nonreflecting combined waveguide junction unless the generator and load are nonreflecting. This is evident if we start with a bilaterally (nonreflecting) matched combined junction and observe that since $\Gamma_B = 0$, the generator does not experience the conjugate matched condition. Hence it is possible to increase the power absorbed by the load by means of adjustment of T_B.

However, when $\Gamma_G = \Gamma_L = 0$, the adjustments of T_A and T_B previously described do result in a combined junction that is nonreflecting. This is evident when it is recalled that for minimum loss $\Gamma_B = \Gamma_G^*$ and $\Gamma_A = \Gamma_L^*$, and if $\Gamma_G = \Gamma_L = 0$, then $\Gamma_A = \Gamma_B = 0$. Thus the intrinsic attenuation described above is equivalent to the intrinsic insertion loss defined by Tomiyasu when $S_{12} = S_{21}$.

In general, one employs lossless tuners as shown in Fig. 1 to obtain a nonreflecting combined junction, and

[4] In practice, this could be done by maintaining the net power input to the network constant by means of a servo system, and adjusting T_A for maximum power absorbed by the load which could conveniently be a detector.

[5] This follows from the definition of efficiency and the scattering equations, recalling that the net power flow across a reference plane in a waveguide is given by

$$P = \frac{|a|^2 - |b|^2}{Z_0},$$

where a and b are the amplitudes of the incident and reflected voltage waves, respectively, and Z_0 is the characteristic impedance.

[6] A graphical construction gives the maxmum efficiency directly from the (measured) locus of input reflection coefficient corresponding to all possible reactive terminations of the two-port. See H. M. Altschuler, "Maximum efficiency of four-terminal networks," Proc. IRE (*Correspondence*), vol. 43, p. 1016; August, 1955.

[7] R. W. Beatty, "Maximum efficiency of a two-arm waveguide junction," IEEE Trans. on Microwave Theory and Techniques (*Correspondence*), vol. 11, p. 94; January, 1963.

it cannot always be obtained simply by connecting appropriate susceptances at the input and output ports as stated by Tomiyasu.[8] This can be seen clearly by inspection of Fig. 2. It is known[1] that any linear, passive, reciprocal two-port can be represented at a single frequency by the circuit of Fig. 2(a), which consists of a bilaterally matched nonreflecting waveguide junction imbedded between two lossless networks, each consisting of a shunt susceptance b and a length l of lossless waveguide.

It is apparent that the addition of tuners having the equivalent circuits shown in Fig. 2(b) will result in a bilaterally matched nonreflecting waveguide junction, since the lengths of line become half wavelength and the susceptances cancel. (The line lengths may be chosen so that integral multiples of half wavelengths are obtained.) This has been shown previously[9,10] for an alternate form of the circuit of Fig. 2(a) by Altschuler.

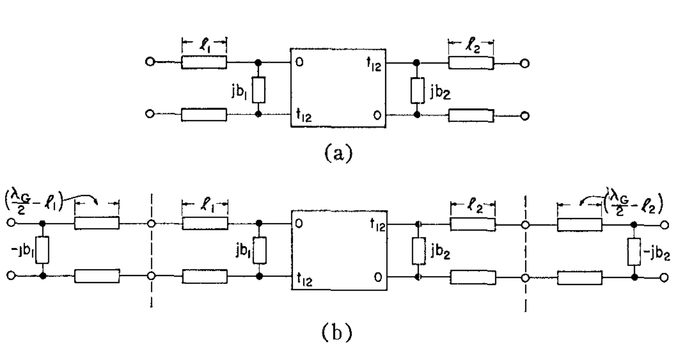

Fig. 2—Equivalent circuit representations. (a) Modified Wheeler network (Tomiyasu). (b) Equivalent circuit of lossless transformers needed to produce bilateral (nonreflective) match.

Expected Benefits of Bilateral Match

As a practical matter, one may be interested in the amount by which the attenuation of a waveguide junction is reduced by tuning for the bilaterally matched nonreflecting condition. Denoting the attenuation of the junction by

$$A = 10 \log_{10} \left[\frac{1}{|S_{21}|^2} \left(\frac{Z_{02}}{Z_{01}} \right) \right], \tag{5}$$

we can form the difference $A - A_I$ and calculate this from the scattering coefficients of the junction, if they are given.

One would ordinarily require three complex coefficients to calculate $A - A_I$, but it is possible to obtain a useful graph from more limited information. For example, if we consider symmetrical attenuators for which we know the attenuation and VSWR, then $|S_{12}S_{21}|$, $|S_{11}|$, and $|S_{22}|$ can be easily determined. Allowing the phases of the scattering coefficients to vary arbitrarily, we can calculate the limits of $A - A_I$ and obtain an idea of how much reduction in attenuation is to be expected from tuning for bilateral match.

The graph resulting from this procedure is shown as Fig. 3. It shows for example that one could not expect much reduction in the attenuation of typical commercially available attenuators by tuning for nonreflection. However, if the mismatch is large (VSWR >2) the expected reduction may be as great as 1 db or more.

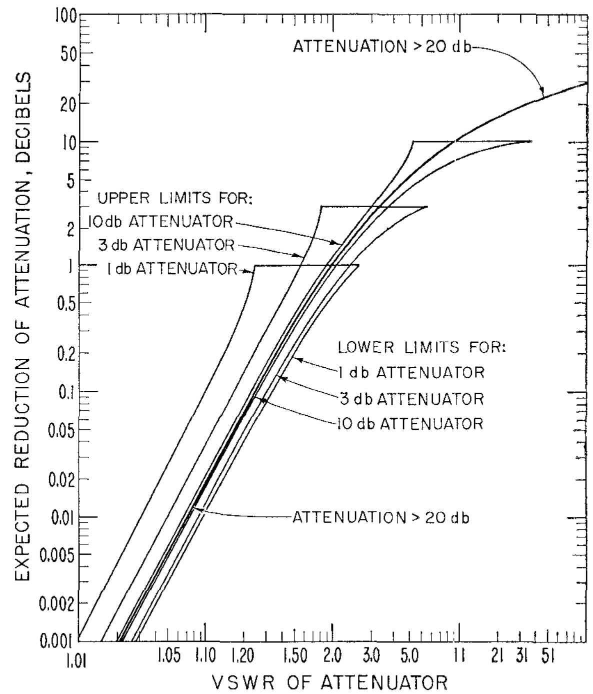

Fig. 3—Limits of expected reduction in attenuation of symmetrical attenuators when tuning for bilateral nonreflecting match.

Experimental Results

It has been assumed that lossless transformers are used when tuning for the bilateral nonreflecting matched condition. However, in practice, the transformers will have some loss and the expected reduction in attenuation will not be completely realized.

In order to obtain some idea of the losses in tuning transformers, experiments were performed with poorly matched coaxial attenuators and the results are shown in Fig. 4. It is apparent that the transformer losses amount to a few tenths of a decibel, and when the expected reduction in attenuation is of the same order of magnitude, one has little or nothing to gain by tuning for bilateral match. In fact one may lose in some cases.

The results would be expected to be better for trans-

[8] Actually Tomiyasu meant to say "suitable shunting susceptances near the ports, including any required line lengths," private communication.

[9] H. M. Altschuler, "A method of measuring dissipative fourpoles based on a modified Wheeler network," IRE Transactions on Microwave Theory and Techniques, vol. 3, pp. 30–36; January, 1955.

[10] H. M. Altschuler and W. K. Kahn, "Nonreciprocal two-ports represented by modified Wheeler networks," IRE Trans. on Microwave Theory and Techniques, vol. MTT-4, pp. 228–233; October, 1956.

PAD	VSWR		Attenuation of Pad	Attenuation of Pad and Tuners	Actual Reduction	Expected Reduction
	Male End	Female End				
A	1.75	1.54	7.368 DB	7.166 DB	.202 DB	0.3-0.9 DB
B	1 14	1 08	6.776 DB	7.079	-.303	.01-.04 DB
C	1.32	1.06	10 392 DB	10.641	-.249	.14-.18 DB

Fig. 4—Expected vs actual reduction in attenuation for coaxial pads measured at 4 Gc.

formers in rectangular waveguide in which there is no dielectric support required for the center conductor. Also it is apparent that tuning for bilateral match can be worthwhile when the degree of mismatch is large (VSWR > 2).

Conclusions

It was noted that "insertion loss" in general depends not only upon the parameters of the network, but also upon the characteristics of the system into which it is inserted. Hence the minimum insertion loss of a network cannot in general be regarded as intrinsic to the network, and the term "intrinsic attenuation" is recommended. Formulas were given to permit calculation of this quantity, given the scattering coefficients of the network. A graph was presented to rapidly estimate the reduction in attenuation to be expected when tuning a symmetrical attenuator for bilateral match, given only the attenuation and VSWR of the attenuator. The significance of tuning for bilateral match was explained, and a method given for obtaining this condition. Experimental results indicated that in the case of mismatched attenuators, little or nothing is to be gained when tuning for bilateral match, and, in some cases, the losses in the tuning transformers may cause a net increase in loss, rather than a reduction. However, in a situation in which the degree of mismatch is large (VSWR > 2), a significant decrease in loss should be obtained when tuning for the bilateral nonreflecting matched condition.

Acknowledgment

Calculations for the graph of Fig. 3 were performed by W. A. Downing, Jr. Measurements for Fig. 4 were performed by W. E. Little.

Analytical Solution to a Waveguide Leaky-Wave Filter Structure*

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Summary—Leaky-wave absorption filters have been found advantageous for the suppression of spurious energy of high-power transmitters. However, although there are experimental data of the properties of several specially constructed leaky-wave filters, there are apparently little data relating the effect upon the attenuation of the filters of varying one or more of the possible parameters of their design. In this paper a waveguide leaky-wave filter structure that retains the basic geometry of waveguide leaky-wave filters is analyzed theoretically over a finite frequency range. The complex propagation constant for the least-attenuated leaky-wave mode is obtained by reducing the fundamental integral equation to a transverse resonance equation and solving the reduced equation. The attenuation constant of the least-attenuated mode is obtained for values of $2a/\lambda$ (*i.e.*, the ratio of waveguide width to one half the free-space wavelength) ranging from 0 to 2. Its dependence on various design parameters of leaky-wave filters, such as main waveguide height, spacing of the coupling slots, width of coupling slots and height of the absorbing waveguides is presented. Good correspondence between theoretically computed curves and experimental data was obtained.

* Received October 1, 1962; revised manuscript received January 18, 1963. The work reported here was sponsored by the Air Force Systems Command, Rome Air Development Center, Griffiss Air Force Base, New York, N. Y., under Contract No. AF 30(602)-2734.
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Introduction

In recent years waveguide structures that support leaky-wave modes[1,2] have been found advantageous for the suppression of spurious energy of high-power transmitters. These structures are generally referred to as leaky-wall or leaky-wave filters.[3–5] They are absorption rather than reflection filters and exhibit the following characteristics:

1) They generally provide high attenuation throughout a wide stop band.

[1] N. Marcuvitz, "On field representations in terms of leaky modes or eigenmodes," IRE Trans. on Antennas and Propagation, vol. AP-4, pp. 192–194; July, 1956.

[2] K. G. Budden, "The Wave-Guide Mode Theory of Wave Propagation," Prentice-Hall, Inc., Englewood Cliffs, N. J., pp. 4, 134; 1961.

[3] V. Met, "Absorptive filters for microwave harmonic power," Proc. IRE, vol. 47, pp. 1762–1769; October, 1959.

[4] V. Price, R. Stone and V. Met, "Harmonic Suppression by Leaky-Wall Waveguide Filter," 1959 IRE Wescon Convention Record, pt. 1, pp. 112–116.

[5] E. G. Cristal, "Some preliminary experimental results on coaxial absorption leaky-wave filters," *4th Ann. Symp. on Radio Frequency Interference*, San Francisco, Calif., June 28–29, 1962.