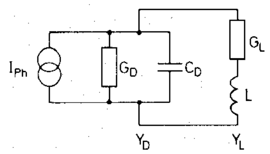

Fig. 1. Lumped model of BARITT oscillator used for locking analysis.

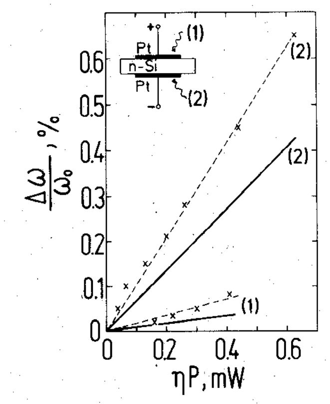

Fig. 2. Locking characteristics of BARITT oscillator. $\omega_o = 3.88 \times 10^9$ s^{-1}, $Q_L = 17$, $P_{HF} = 4.3$ mW, $G_L = 0.37$ mS. (1) Optical hole injection, $\tau_p = 1.45$ ns. (2) Optical electron injection, $\tau_n = 0.54$ ns. × × × × ×: measured; ——: calculated.

where ω_{Ph} denotes the modulation frequency of the optical signal, τ the transit time of the optically injected holes or electrons, respectively, P the total sideband power of the modulated optical signal, λ the vacuum wavelength of the light, and η a correction factor taking into account the quantum efficiency and the losses due to reflection at and absorption in the metallic contact. If neither the velocity of holes nor the velocity of electrons is saturated within the BARITT device, the transit time τ_n of electrons is shorter than the transit time τ_p of holes, hence the locking bandwidth in case of optical electron injection should be larger than in case of optical hole injection.

In the locking characteristics of Fig. 2, experimental and theoretical results are compared for optical hole and electron injection. As can be seen, the measured locking bandwidth increases linearly with optical power, and the locking bandwidth in case of electron injection is much larger than in the case of hole injection, as predicted by theory. Intensity modulated laser light ($\lambda = 633$ nm) was used as the optical locking signal. The loaded Q factor was measured with conventional electronic injection locking. The theoretical characteristics were calculated using the measured oscillator and circuit parameters, the numerically calculated small-signal admittance Y_D, and the transit times τ_p and τ_n.

As a result, the simple theory offers reasonable correlation as compared with the experimental results. However, the measured locking bandwidths are approximately 50 percent larger than the calculated bandwidths. This may be traced back to the fact that the simple theory does not allow for an optical modulation of the device admittance Y_D and its influence upon the locking characteristics. The locking bandwidth of about 0.6 percent as obtained with BARITT oscillators, is similar to the bandwidths predicted for IMPATT oscillators [1] and for transistor oscillators [2], [3]. Experimentally, it has been shown that optical injection locking of BARITT oscillators is possible. A locking bandwidth of 0.6 percent is obtained which may be viewed as a useful value. There is reasonable agreement between measurement and theory. In the higher microwave range, it is recommended to use light modulated by a subharmonic of the oscillator frequency as the optical injection signal because, at present, it might be difficult to modulate laser light using high microwave frequencies.

References

[1] J. R. Forrest and A. J. Seeds, "Optical injection locking of IMPATT oscillators," *Electron. Lett.*, vol. 14, pp. 626–627, Sept. 1978.
[2] H. W. Yen and M. K. Barnoski, "Optical injection locking and switching of transistor oscillators," *Appl. Phys. Lett.*, vol. 32, pp. 182–184, Feb. 1978.
[3] A. A. de Salles and J. R. Forrest, "Initial observations of optical injection locking of GaAs MESFET oscillators," *Appl. Phys. Lett.*, vol. 38, pp. 392–394, Mar. 1981.
[4] R. Adler, "A study of locking phenomena in oscillators," *Proc. IRE*, vol. 34, pp. 351–357, June 1946.
[5] W. W. Gärtner, "Depletion-layer photoeffects in semiconductors," *Phys. Rev.*, vol. 116, pp. 84–87, Oct. 1959.

Comments on "A Rigorous Technique for Measuring the Scattering Matrix of a Multiport Device with a Two-Port Network Analyzer"

H. DROPKIN, SENIOR MEMBER, IEEE

The recent article[1] by Tippet and Speciale uses the matrix formulation of the generalized scattering parameter renormalization transformation in the form

$$S' = (I-S)^{-1}(S-\Gamma)(I-S\Gamma)^{-1}(I-S). \qquad (1)$$

Here S is the $N \times N$ scattering matrix of an N port with port line impedances ζ, S' is the transformed scattering matrix when the port impedances are altered to Z, and Γ is the diagonal matrix of reflection coefficients of Z as seen from line impedances ζ, and I is the identity matrix.

Equation (1) can be simplified as follows:

$$S = (I-S)^{-1}S(I-S) \qquad (2)$$

$$S' - S = (I-S)^{-1}\left[(S-\Gamma)(I-S\Gamma)^{-1} - S\right](I-S). \qquad (3)$$

The bracketed term is

$$\begin{aligned}(S-\Gamma)(I-S\Gamma)^{-1} &- S(I-S\Gamma)(I-S\Gamma)^{-1}\\ &= [(S-\Gamma)-S(I-S\Gamma)](I-S\Gamma)^{-1}\\ &= -(I-S^2)\Gamma(I-S\Gamma)^{-1}.\end{aligned}$$

Note the cancellation of the individual S terms. Since $(I-S^2) = (I-S)(I+S)$, the prefactor in (3) also cancels, with the final result

$$S' = S - (I+S)\Gamma(I-S\Gamma)^{-1}(I-S). \qquad (4)$$

Equation 4 has reduced the number of matrix inversions needed from 2 to 1. Also, S' is now obtained by an additive correction to

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The author is with the U.S. Army Research and Development Command, Harry Diamond Laboratories, Adelphi, MD 20783.

[1] J. C. Tippet and R. A. Speciale, *IEEE Trans. Microwave Theory Tech.*, vol. MTT 30, pp. 661–666, May 1982.

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be applied to S, and these corrections contain Γ as an explicit factor. When Γ is small (small line impedance changes), a good first order approximation is

$$S' = S - (I+S)\Gamma(I-S) \tag{5}$$

which has no matrix inversions.

In some cases the inverse required by (1) but not by (4) may not even exist. A simple example is given by a matched lossless two port for which

$$S = \begin{pmatrix} 0 & e^{i\theta} \\ e^{i\theta} & 0 \end{pmatrix}$$

and for which $(I - S)$ has determinant 0 for some values of θ. Clearly, less extreme cases can be numerically difficult, especially since the components of S are to be experimentally measured.

This form of the transformation should require less computation and have improved accuracy.

Reply[2] by John C. Tippet and Ross A. Speciale[3]

The authors of the original paper[1] would like to warmly commend Dr. H. Dropkin for constructively contributing a further simplification of the generalized scattering matrix renormalization transform, and would like to take this opportunity to add some related comments.

First, it is interesting to notice the alternative formulation of the generalized transform, given in the original 1980 IEEE-CAS symposium paper [2]

$$S' = (I-\Gamma)^{-1}(S-\Gamma)(I-\Gamma S)^{-1}(I-\Gamma). \tag{1}$$

This alternative form of the transform was obtained as an application of the projective matrix transformation

$$S' = (T_1 \cdot S + T_2)(T_3 \cdot S + T_4)^{-1} \tag{2}$$

first introduced in 1977 [3] and later generalized in 1981 [4]. The form (1) has been successfully verified analytically in the 2×2 case of two-port networks [5] and it is, therefore, believed to be totally equivalent to the original form given in [1]

$$S' = (I-S)^{-1}(S-\Gamma)(I-S\Gamma)^{-1}(I-S) \tag{3}$$

and to the simplified formulation found by Dropkin

$$S' = S - (I+S)\Gamma(I-S\Gamma)^{-1}(I-S). \tag{4}$$

We have been so far unable to prove analytically the equivalence of the alternative form (1) the original form (3), in the general case. It is, however, interesting to notice that, by exploiting the identity

$$\Gamma = (I-\Gamma)^{-1}\Gamma(I-\Gamma) \tag{5}$$

and proceeding in a way similar to that suggested by Dropkin, it is possible to simplify the form (1) to

$$S' = (I+\Gamma)S(I-\Gamma S)^{-1}(I-\Gamma) - \Gamma. \tag{6}$$

This second simplified form (6) also requires only one matrix inversion and the matrix to be inverted, $I - \Gamma S$, is in general different from the matrix $I - S\Gamma$ that must be inverted when using the Dropkin form (4). These matrices are either both singular or both nonsingular at the same time. Furthermore, it should be easier to prove the mutual equivalence of the two simplified forms (4) and (6) analytically. While attempting this proof, we would like to propose it as a challenge to the adventurous reader. Another aspect of the generalized renormalization transform that we have not been able to prove analytically relates to what may be called the inverse transform, expressing the matrix S as function of the matrix S'. Considering the definition of the reflection-coefficient-matrix Γ we would expect to be able to invert any of the forms (1), (3), (4), and (6) by simply exchanging the matrices S and S', while at the same time changing the sign of the matrix Γ. In this connection, it is easy to see that the form (6) may be solved for the matrix S, obtaining

$$S = (I-\Gamma)(I+S'\Gamma)^{-1}(S'+\Gamma)(I-\Gamma)^{-1} \tag{7}$$

and thus this inverted form may be proved to be equivalent to

$$S = (I-\Gamma)S'(I+\Gamma S')^{-1}(I+\Gamma)+\Gamma \tag{8}$$

and should also be equivalent to

$$S = S' + (I+S')\Gamma(I+S'\Gamma)^{-1}(I-S'). \tag{9}$$

These last two forms have been obtained from (6) and (4), respectively, by mutually exchanging the matrices S and S' and changing the sign of the matrix Γ, as indicated before.

Finally, we would like to point out that the success of our experimental verification of the renormalization transform is clearly a consequence of the fact that our transform was derived for scattering matrices defined on the basis of voltage waves (or traveling waves) rather than for power-wave scattering matrices. This is the kind of scattering matrix obtained from any calibrated automated network analyzer. In these respects, we would like to express our total agreement with Professor Wood's statement [6] declaring the impossibility of performing power-wave scattering measurements on any known form of ANA. Indeed, all known forms of ANA use voltage sensors to monitor the incident, reflected, and transmitted waves and could not possibly sense power-waves under practical conditions, where the ANA measurement-port impedances, as seen from the device under test, are both complex and frequency-dependent. A generally unrecognized aspect of ANA calibration methods is that these have the capability of eliminating measurement-port mismatch-errors, because of including implied voltage-wave-scattering-matrix renormalizations to the nominal wave-impedance of at least one of the calibration standards.

It would appear that the concept of power-wave scattering matrix has outlived its usefulness, especially considering the fact that the net active power flow entering a network through any given cluster of n ports may be expressed by means of the Hermitian form

$$P_{in} = \begin{vmatrix} b_i \\ \hline a_i \end{vmatrix}^* \cdot \begin{vmatrix} -\left[(Z_i^*)^{-1} + Z_i^{-1}\right] & -\left[(Z_i^*)^{-1} - Z_i^{-1}\right] \\ \hline (Z_i^*)^{-1} - Z_i^{-1} & (Z_i^*)^{-1} + Z_i^{-1} \end{vmatrix} \cdot \begin{vmatrix} b_i \\ \hline a_i \end{vmatrix} \tag{10}$$

Where the a_i's and b_i's $(i = 1, \cdots, n)$ are the incident and emerging voltage-waves, respectively, and Z_i is the wave-normalization impedance matrix $(Z_i^* = \bar{Z}')$.

[2]Manuscript received July 26, 1982.

[3]The authors are with TRW Defense and Space Systems Group, Redondo Beach, CA 90278

REFERENCES

[1] R A Speciale and J C Tippet, "A rigorous technique for measuring the scattering matrix of a multi-port device with a two-port network analyzer," *IEEE Trans. Microwave Theory Tech.*, vol MTT-30, pp 661–666, May 1982

[2] R A Speciale, "Derivation of the generalized scattering parameter renormalization transformation," in *Proc. 1980 IEEE CAS Int. Symp Circuits Syst.* (Houston, TX), Apr. 28–30, 1980, pp 166–169.

[3] R A. Speciale, "A generalization of the TSD network analyzer calibration

procedure, covering n-port measurements, affected by leakage errors," *IEEE Trans. Microwave Theory Tech* vol. MTT-25, pp. 1100–1115, Dec. 1977

[4] R. A. Speciale, "Projective matrix transformations in microwave network theory," in *1981 IEEE MTT-S Int. Microwave Symp. Dig. Tech Papers*, (Los Angeles, CA), June 15–19, 1981, pp 510–512.

[5] R. A Speciale, "Derivation of the generalized scattering parameter renormalization transformation," presented at Fall 1979 14th Workshop Meet. Automated Radio Frequency Tech. Group IEEE-MTT (ARFTG), (Los Angeles, CA), Nov. 11–12, 1979

[6] D Woods, "Multiport-network analysis by matrix renormalization employing voltage-wave S-parameters with complex normalization," *Proc. Inst. Elec. Eng*, vol. 124, pp. 198–204, Mar. 1977.

Comment on "The Exact Noise Figure of Amplifiers with Parallel Feedback and Lossy Matching Circuit"

JAKOB ENGBERG

I found the short paper of K. B. Niclas "The Exact Noise Figure of Amplifiers with Parallel Feedback and Lossy Matching Circuits"[1] very interesting, but, except for the approximated formulas, it contained only little new information, since my paper "Simultaneous Input Power Match and Noise Optimization using Feedback" [2] included most of the formulas. In that paper, I have developed a general form of noise parameters of a three-port with combinations of parallel and series feedback (or imbedding) elements. These formulas include both the formula of Niclas and of Iversen [1, [9]]. The formulas have been used in the lossless case to develop an amplifier with simultaneous noise and power input match.

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The author is with the Institut for Elektroniske Systemer, Aalborg Universitetscenter, 9000 Aalborg, Denmark.

[1]K. B. Niclas, *IEEE Microwave Theory Tech.*, vol. MTT-30, pp. 832–834, May 1982.

Reply[2] by Karl B. Niclas[3]

While I was aware of the existence of Mr. Engberg's paper, "Simultaneous Input Power Match and Noise Optimization using Feedback" [2], I was not able to secure a copy and consequently was unaware of its contents until his comments arrived. However, after careful study of his publication, I have to disagree with his conclusion that the information in my short paper contains only little new information outside of the approximation formulas. It must have escaped Mr. Engberg that his formulas for the equivalent noise parameters of the parallel feedback amplifier (3) in [2] deviate from mine (4) in [1] and only for the special case of $Y_B = 0$ ($Y_{FB} = 0$) was I able to find agreement for the parameters R'_n and Y'_γ (R^I_n and Y^I_{cor}). Since no derivations are presented in Mr. Engberg's paper, I am not in a position to explain the discrepancies. It should be noted, however, that results calculated with formulas given in my paper are in perfect agreement with those computed with the aid of Compact which is based on Hillbrand's and Russer's correlation matrix [3], [4].

References

[1] Karl B. Niclas, "The exact noise figure of amplifiers with parallel feedback and lossy matching circuits," *IEEE Trans. Microwave Theory Tech* vol. MTT-30, pp. 832–34, May 1982.

[2] Jakob Engberg, "Simultaneous input power match and noise optimization using feedback," in *Conf. Proc. 4th European Microwave Conf.* (Montreux, Switzerland), September 1974, pp. 385–389.

[3] H. Hillbrand and P. H. Russer, "An efficient method to computer-aided noise analysis of linear amplifier networks," *IEEE Trans. Circuits Syst.*, vol. CAS-23, pp. 235–238, Apr. 1976

[4] Private conversation with C. Holmes of Compact Engineering, Palo Alto, CA.

[2]Manuscript received July 25, 1982

[3]The author is with the Devices Group, Watkins-Johnson Company, 3333 Hillview Ave., Stanford Industrial Park, Palo Alto, CA 94304