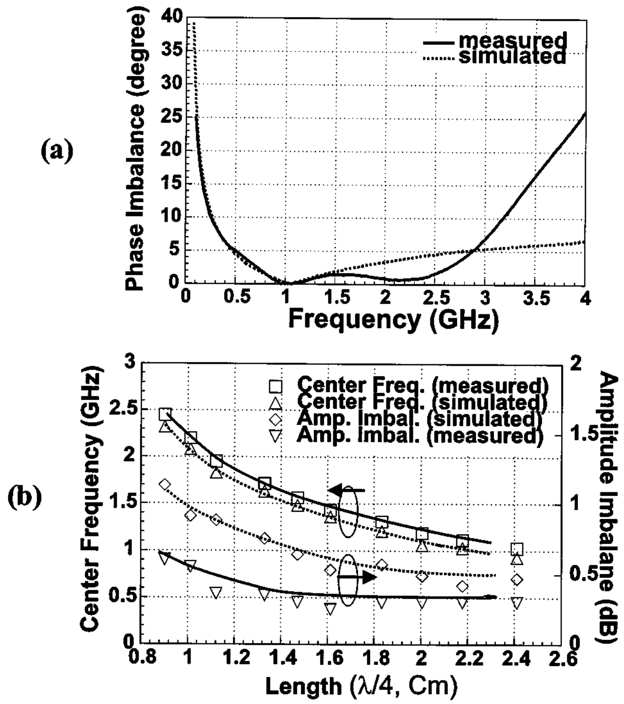

Fig. 4. Comparison of: (a) simulated phase deviation with measurement for the spiral transmission-line balun and (b) simulation with measurement in terms of frequency of bandwidth and amplitude imbalance for 50-μm-conductor-width baluns built on high-resistivity silicon substrates.

beyond the second order slightly improve the accuracy with the price of high model–circuit complexity.

IV. Conclusions

In this paper, we have presented the first- and second-order equivalent lumped-element models for spiral transmission-line baluns. The second-order and higher order models are synthesized from the first-order model. Since the model parameters are based on lumped values extracted from the physical structures of the devices, the models are useful for predicting the major characteristics from the design parameters of baluns. In comparison with the measurement results, the second-order model captures all important features of the behaviors of the spiral couplers and transmission-line baluns within 10% error.

References

[1] N. Marchand, "Transmission line conversion transformers," *Electronics*, vol. 17, pp. 142–145, Dec. 1944.

[2] W. R. Brinlee, A. M. Pavio, and K. R. Varian, "A novel planar double balanced 6–18 GHz MMIC mixer," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1994, pp. 9–12.

[3] T. Chen, K. W. Chang, S. B. Bui, H. Wang, G. Samuel, L. C. T. Lui, T. S. Lin, and W. S. Titus, "Broad-band monolithic passive baluns and monolithic double-balanced mixer," *IEEE Trans. Microwave Theory Tech.*, vol. 39, pp. 1980–1986, Dec. 1991.

[4] M. Engels and R. H. Jansen, "Design of integrated compensated baluns," *Microwave Opt. Technol. Lett.*, vol. 14, no. 2, pp. 75–81, 1997.

[5] W. K. Roberts, "A new wide-band balun," *Proc. IRE*, vol. 45, pp. 1628–1631, Dec. 1957.

[6] Y. J. Yoon, Y. Lu, R. C. Frye, and P. R. Smith, "A silicon monolithic spiral transmission line balun with symmetrical design," *IEEE Electron. Device Lett.*, vol. 20, pp. 182–184, Apr. 1999.

[7] ——, "Spiral transmission line baluns for RF multi-chip module packages," *IEEE Trans. Comp., Packag., Manufact. Technol.*, vol. 22, pp. 332–336, Aug. 1999.

[8] E. Pettenpaul, H. Kapusta, A. Weisgerber, H. Mampe, J. Lugunsland, and I. Wolff, "CAD models of lumped elements on GaAs up to 18 GHz," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 294–304, Feb. 1988.

[9] *Transmission Line and Crosstalk Products User's Manual, XNS-TLC-XFX*, Mentor Graphics, Wilsonville, OR, 1996.

[10] E. Chen and S. Y. Chou, "Characteristics of coplanar transmission lines on multilayer substrates: Modeling and experiments," *IEEE Trans. Microwave Theory Tech.*, vol. 45, pp. 939–945, June 1997.

[11] K. C. Gupta, R. Grag, I. Bahl, and P. Bhartia, *Microstrip Lines and Slotlines*, 2nd ed. Norwood, MA: Artech House, 1996, ch. 7.

[12] *Libra Reference Manual in HP EEsof Series IV*, Hewlett-Packard, Santa Rosa, CA, 1996.

[13] Y. J. Yoon, Y. Lu, R. C. Frye, M. Lau, P. R. Smith, L. Alquist, and D. Kossives, "Design and characteristics of multilayer spiral transmission line baluns," *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 1841–1847, Sept. 1999.

Transmission-Line Stabilized Monolithic Oscillators

Kian Sen Ang, Michael J. Underhill, and Ian D. Robertson

***Abstract*—A novel technique for frequency stabilization and phase-noise reduction of monolithic oscillators is presented in this paper. It employs simple transmission-line resonators, which are many wavelengths long to increase the oscillator quality factor. Monolithic oscillators at 20 and 40 GHz are realized for the application of this technique. Phase noise reduction of more than 20 dB was achieved for both oscillators. The single-sideband phase noise obtained was −100 dBc/Hz at 100-KHz offset for the 20-GHz oscillator and −90 dBc/Hz at 1-MHz offset for the 40-GHz oscillator. The approach is implemented by using readily available transmission lines, which are open- or short-circuited at one end and connected to the monolithic-microwave integrated-circuit (MMIC) oscillator at the other end. Thus, it presents significant potential in the development of low-cost MMIC oscillators with enhanced noise performance.**

***Index Terms*—Frequency stability, MMIC oscillators, phase noise, transmission-line resonators.**

I. Introduction

In recent years, an increasing number of oscillator circuits has been implemented using monolithic technology due to the overwhelming advantages of size, reliability, and cost [1]. These oscillators, however, generally have poor phase-noise performance as the monolithic circuitry has relatively low quality factor (Q). To meet the low noise requirements of practical systems, some form of an oscillation stabilizing scheme has to be employed.

The most common frequency stabilization and phase-noise reduction techniques are the use of dielectric resonator oscillators (DROs) [2], [3] and phase-locked oscillators (PLOs) [4], [5]. Due to the extraordinarily high-Q, DROs offer excellent low noise performance. However, they require accurate dimensions of the dielectric puck and its precise placement on the monolithic or package substrate, as these parameters determine the oscillating frequency and stability. This demands careful

Manuscript received October 4, 1999. This work was supported in part by the Engineering and Physical Sciences Research Council and in part by Prof. K. K. Cheng, Chinese University of Hong Kong. The work of K. S. Ang was supported in part by the Defence Science Organization National Laboratories.

The authors are with the Microwave and Systems Research Group, School of Electronic Engineering, Information Technology and Mathematics, University of Surrey, Guildford, Surrey GU2 7XH, U.K. (e-mail: k.ang@eim.surrey.ac.uk).

Publisher Item Identifier S 0018-9480(01)01086-9.

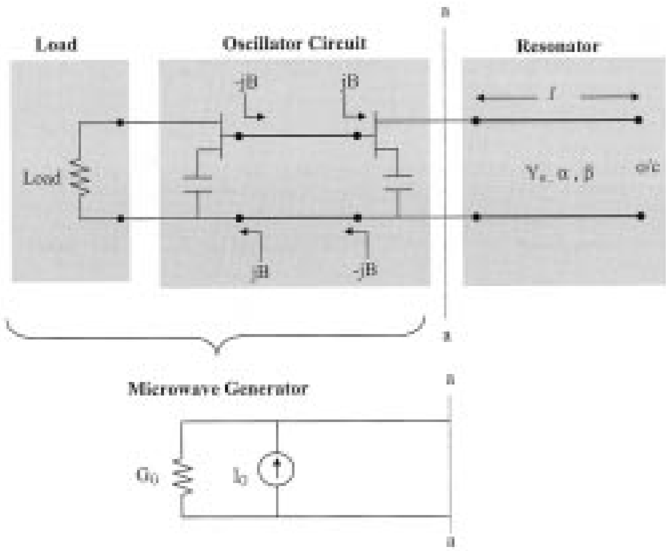

Fig. 1. Block diagram of a microwave oscillator. The oscillator circuit employs a balanced oscillator, while the resonator employs open-circuit transmission-line resonators. The oscillator and load constitute a microwave generator at the resonator plane a–a.

post-fabrication attention, resulting in high manufacturing cost and low repeatability. The alternative approach of using PLOs requires many components such as voltage-controlled oscillators, frequency dividers, and phase/frequency comparators. Furthermore, additional frequency multipliers and amplifiers are required due to the limited operating frequency of frequency dividers. These requirements result in a complex and high-cost multichip scheme at millimeter-wave frequencies.

In this paper, a stabilizing technique using simple transmission-line resonators is presented. It does not require precise placement of the resonator, as the transmission-line resonator is connected directly to the monolithic oscillator. The oscillation frequency is determined by the resonator length, which can be readily varied during manufacture to obtain the desired frequency. Electronic control of the resonator's electrical length can also be implemented by incorporating varactor diodes at the end of the transmission line, either on-chip within the monolithic oscillator or off-chip at the other end of the transmission line [6]. This provides additional frequency tuning capability to compensate for monolithic-microwave integrated-cirucit (MMIC) process variations and frequency drifts due to temperature variations. This results in a simple and low-cost stabilizing scheme. A simplified derivation of the improvement in oscillator Q and measurement results of K- and Ka-bands monolithic oscillators are presented to demonstrate the technique.

II. Transmission-Line Resonators

The most effective way of improving an oscillator's noise performance is to increase its Q factor, as it determines the oscillator's operating parameters such as phase noise, frequency stability, and pulling figure. In this section, the feasibility of using long transmission-line resonators to increase the loaded Q of oscillator circuits is investigated.

Fig. 1 shows a functional diagram of a microwave oscillator. Maintaining the resonator plane a–a as a reference, the oscillator circuit and load constitute a microwave generator. A constant current source I_G in parallel with source admittance G_G is used to represent the generator. The resonator is realized using a transmission line with characteristic admittance Y_0 and electrical length l. Open-circuited transmission-line resonators will be considered, although equivalent results can be obtained with short-circuited resonators. The input admittance of the transmission-line resonator can be expressed as

$$Y_{\text{in}} = G_{\text{in}} + jB_{\text{in}} \tag{1}$$

where

$$G_{\text{in}} = Y_0 \frac{\sinh(\alpha\ell)\cosh(\alpha\ell)}{\sinh^2(\alpha\ell) + \cos^2(\beta\ell)} \tag{2a}$$

$$B_{\text{in}} = Y_0 \frac{\sin(\beta\ell)\cos(\beta\ell)}{\sinh^2(\alpha\ell) + \cos^2(\beta\ell)} \tag{2b}$$

α and β are the attenuation and phase constants of the transmission line. At frequencies where l is an integer number of half-wavelengths

$$\beta\ell = N\pi \tag{3}$$

where N is an integer. The input admittance is then given by

$$Y_{\text{in}} = G_{\text{in}} = Y_0 \tanh(\alpha\ell). \tag{4}$$

At frequencies in the vicinity of $\beta l = N\pi$, the input characteristics of the transmission line can be modeled by a parallel resonant circuit. The loaded Q factor of the resonant circuit can be evaluated as [7]

$$Q_L = \frac{\omega_r}{2\left(G_G + G_{\text{in}}\big|_{\omega=\omega_r}\right)} \frac{dB_{\text{in}}}{d\omega}\Bigg|_{\omega=\omega_r} \tag{5a}$$

$$= \frac{Y_0 N\pi}{2\left(G_G + Y_0 \tanh\left(\frac{\alpha N\lambda_r}{2}\right)\right)\left(\cosh^2\left(\frac{\alpha N\lambda_r}{2}\right)\right)} \tag{5b}$$

where λ_r is the resonant wavelength.

From (5b), it appears that, by increasing the length of the open-circuited transmission line, Q_L can be increased. This is valid if the transmission line is of sufficiently low loss such that the terms of $(\alpha N\lambda_r)$ are negligible. As Q_L increases with increasing length, the associated line losses will become increasingly significant. This will result in decreasing improvement with increasing length. Eventually, as the line losses become excessive, Q_L will start to decrease.

Another effect of increasing line length is that the transmission line will have many closely spaced resonant frequencies. Thus, the circuit will have a spurious response close to the desired resonant frequency, which is unacceptable in many applications. Therefore, for the approach to be feasible, the oscillator circuit must retain some selectivity, other than that provided by the transmission line. This is to ensure oscillation occurs only at a single mode, whereby other alternative modes are sufficiently suppressed once a single oscillation mode is excited [8]. In addition, it must be able to operate under dual loading conditions, being connected both to an external load and a transmission-line resonator. This requires two output ports with sufficient isolation. If these oscillator requirements are met and the transmission-line resonator is sufficiently low loss, increasing the resonator length is an effective way to increase Q_L and reduce phase noise. A simple oscillator circuit fulfilling these requirements is the balanced oscillator circuit [9].

III. Oscillator Circuit

The schematic diagram of the oscillator circuit is shown within the oscillator circuit block in Fig. 1. It consists of two FETs whose gates are interconnected by a transmission line. Capacitive source feedback is applied to both FETs to generate a negative conductance at the desired oscillation frequency. Looking into the gates of each device is a positive susceptance jB. The length of the transmission line is chosen

TABLE I
MEASURED PARAMETERS OF THE THREE COAXIAL CABLES

Resonator	Cable Length (cm)	No. of half-wavelengths, N		Insertion Loss (dB)	
		20GHz	43GHz	20GHz	43GHz
R1	≅ 10	23	49	0.35	0.7
R2	≅ 50	89	191	1.7	3.2
R3	≅ 100	171	368	3.2	6.5

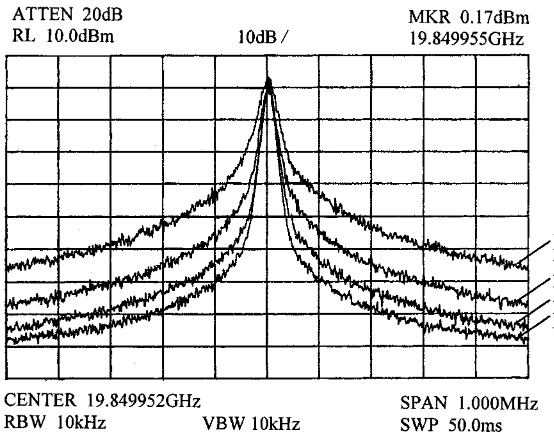

Fig. 2. Frequency spectrums of K-band oscillator with different resonators: $R0$: no external resonator. $R1$: open-circuit resonator of length ≅10 cm. $R2$: open-circuit resonator of length ≅50 cm. $R3$: open-circuit resonator of length ≅100 cm.

such that the two devices resonate one another. This is achieved by selecting the length so that each devices susceptance is transformed to a susceptance with the same magnitude, but opposite in sign. Thus, the device susceptances and the transmission line form a resonant circuit. This provides the required frequency selectivity to permit oscillation at only one of the several closely spaced frequencies where the oscillation condition is satisfied in the external resonator. The oscillator outputs for the external load and transmission-line resonator connections are taken from the drains of the two FETs, providing the required output port isolation.

IV. EXPERIMENTAL RESULTS

Previously designed monolithic oscillators at K- and Ka-bands [9] were used to evaluate the technique experimentally. The oscillator chips were measured on a Cascade Microtech probestation. All RF connections to the chips were made using 200-μm-pitch coplanar probes while dc biasing was applied though probe needles.

The K-band oscillator was first measured by connecting one of the output ports to an HP8563E spectrum analyzer, leaving the other output port open circuited. Stable oscillations at around 20 GHz with 0-dBm output power were observed at $V\text{gs} = 0$ V and $V\text{ds} = 3$ V. Without altering the biasing conditions or the analyzer settings, except the center frequency, the oscillator was then measured with open-circuited coaxial cables connected to the other output port. Three coaxial cables of different lengths were used and their measured parameters are given in Table I.

Fig. 2 shows the measured spectra obtained for the four different resonator loading conditions. The displayed power levels include probe and cable losses, which are about 4 dB in total. Similar spectra were observed when the transmission lines were terminated in a short circuit instead of an open circuit. Spurious oscillations at unwanted signal frequencies were not observed in any of the cases.

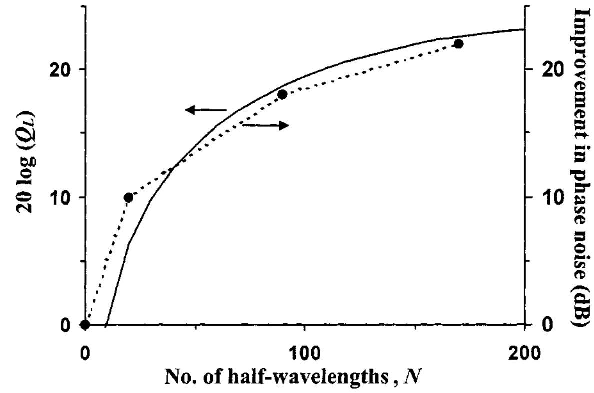

Fig. 3. Improvement in phase noise and $20\log(Q_L)$ versus N for the K-band oscillator.

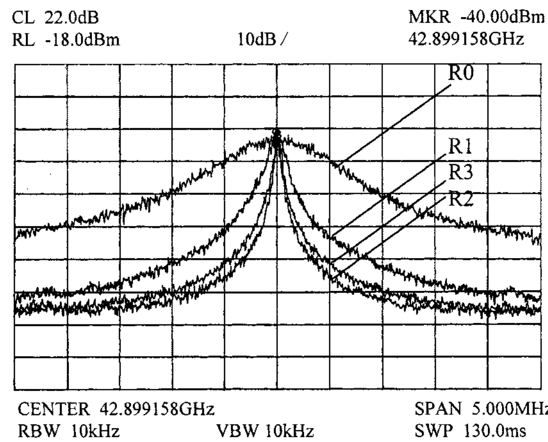

Fig. 4. Frequency spectrums of Ka-band oscillator with different resonators: $R0$: no external resonator. $R1$: open-circuit resonator of length ≅10 cm. $R2$: open-circuit resonator of length ≅50 cm. $R3$: open-circuit resonator of length ≅100 cm.

Fig. 2 depicts a significant 10-dB reduction in phase noise with the connection of a short transmission-line resonator $R1$. Increasing the resonator length by five times, using $R2$, results in a further improvement of about 8 dB. Finally, using resonator $R3$, which is twice the length of $R2$, results in a further 3-dB improvement. The measured phase noise for $R3$ at 100-kHz offset is about -100 dBc/Hz. This trend of decreasing phase-noise improvement with increasing length is depicted in Fig. 3. The resonator length is plotted in terms of N, the number of half-wavelengths. Also plotted is the increase in Q_L as the resonator length increases, given by (5b). Neglecting the effect of nonlinear reactances, phase noise near the carrier is inversely proportional to Q_L^2 [10]. Therefore, $20\log(Q_L)$ is plotted in Fig. 3 for direct comparison with improvement in phase noise. The two curves are in general agreement. Therefore, it is evident that increasing resonator length effectively increases the oscillator Q_L and equivalently reduces phase noise.

The Ka-band oscillator was measured on the spectrum analyzer using external mixers. The same coaxial cables given in Table I were used. Fig. 4 shows the frequency spectra obtained under the different resonator loading conditions, with $V\text{gs} = 0$ V and $V\text{ds} = 2.0$ V. Note that the displayed power level includes probe and cable losses, which are about 7 dB, as well as a 40-dB attenuator, which was included at the input of the external mixer. Fig. 4 depicts a dramatic 20-dB decrease in phase noise with the addition of resonator $R1$. The lowest phase noise

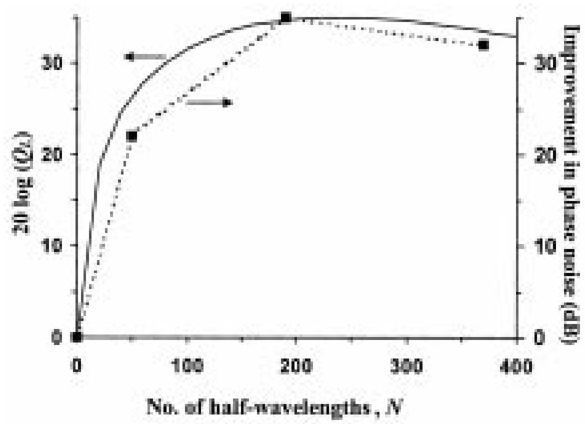

Fig. 5. Improvement in phase noise and $20\log(Q_L)$ versus N for the Ka-band oscillator.

TABLE II
MEASURED FREQUENCY PUSHING OF THE OSCILLATORS UNDER DIFFERENT RESONATOR TERMINATING CONDITIONS

Resonator terminating conditions	Frequency pushing of drain bias (MHz / V)	
	K-band oscillator	Ka-band oscillator
No resonator	217	295
Resonator R1	41	47
Resonator R2	13	26
Resonator R3	5	30

is obtained using resonator $R2$, with −90 dBc/Hz at 1-MHz offset. Increasing the resonator length to $R3$ actually causes 3-dB degradation in phase noise. Fig. 5 shows the measured phase noise improvement and the predicted increase in Q_L as a function of resonator length. The curve of $20\log(Q_L)$ predicts a maximum phase-noise improvement of 35 dB with resonator length of about 200 half-wavelengths. Beyond which, the associated line losses become increasing significant such that Q_L starts to decrease and phase noise degrades with increasing length. This trend is in agreement with the measured results.

In addition to phase-noise reduction, an increase in frequency stability was also observed with increasing resonator length. As an indication of frequency stability, frequency pushing of the two oscillators under different resonator terminating conditions is measured and given in Table II. With the addition of external resonators, a dramatic decrease in frequency pushing was observed for both oscillators.

V. CONCLUSION

This paper has demonstrated that oscillator frequency stabilization and phase-noise reduction can be achieved using simple transmission-line resonators. A simplified derivation of the improvement in phase noise has been verified through measured results of K- and Ka-bands monolithic oscillators. More than 20-dB improvement in phase noise was obtained for both oscillators. These results are obtained through the use of widely available coaxial cables, which may seem impracticably long. The actual length required, however, depend on the frequency of operation, loss characteristics, and dielectric constant of the transmission line. The wide availability of high-quality flexible low-loss cables, which can be compactly coiled, makes this technique a viable option for high-frequency applications. Therefore, the technique offers great potential in the development of low-cost MMIC oscillators with significantly improved noise performance.

ACKNOWLEDGMENT

The authors would like to thank Prof. C. Aitchison, University of Surrey, Guildford, Surrey, U.K., for many valuable discussions.

REFERENCES

[1] I. D. Robertson, Ed., *MMIC Design*. ser. IEE Circuits Syst. 7. London, U.K.: IEE, 1995, pp. 337–370.
[2] P. G. Wilson, "Monolithic 38 GHz dielectric resonator oscillator," in *IEEE Int. Microwave Symp. Dig.*, 1991, pp. 831–834.
[3] M. Funabashi *et al.*, "V-band AlGaAs/InGaAs heterojunction FET MMIC dielectric resonator circuit," in *IEEE GaAs IC Symp. Dig.*, 1994, pp. 30–33.
[4] M. Madihian and K. Honjo, "GaAs monolithic IC's for an X-band PLL-stabilized local source," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-34, pp. 707–713, June 1986.
[5] A. Kanda, T. Hirota, H. Okazaki, and M. Nakamae, "An MMIC chip set for V-band phase-locked local oscillator," in *IEEE GaAs IC Symp. Dig.*, 1995, pp. 259–262.
[6] K. K. M. Cheng and J. K. A. Everard, "Novel varactor tuned transmission line resonator," *Electron. Lett.*, vol. 25, no. 17, pp. 1164–1165, Aug. 1989.
[7] P. A. Rizzi, *Microwave Engineering*. Englewood Cliffs, NJ: Prentice-Hall, 1988, pp. 411–428.
[8] M. J. Underhill, "Fundamentals of oscillator performance," *Electron. Commun. Eng. J.*, vol. 44, pp. 185–193, Aug. 1992.
[9] K. S. Ang, M. J. Underhill, and I. D. Robertson, "Balanced monolithic oscillators at K- and Ka-band," *IEEE Trans. Microwave Theory Tech.*, vol. 48, pp. 187–193, Feb. 2000.
[10] D. B. Leeson, "A simple model of feedback oscillator noise spectrum," *Proc. IEEE*, vol. 54, pp. 329–330, Feb. 1966.

ALPS—A New Fast Frequency-Sweep Procedure for Microwave Devices

Din-Kow Sun, Jin-Fa Lee, and Zoltan Cendes

***Abstract*—The discretization of Maxwell equations results in a polynomial matrix equation in frequency. In this paper, we present a robust and efficient algorithm for solving the polynomial matrix equation. To solve this equation for a broad bandwidth, one previously performs a discrete frequency sweep where the resulting matrix needs to be inverted at numerous frequencies, while current procedure requires only one matrix inversion. Speed improvements compared to the discrete sweep range from 10 to 100 times, depending on number of resonance peaks encountered.**

***Index Terms*—Lanczos algorithm, Maxwell's equations, reduced-order system, transfinite-element method.**

I. INTRODUCTION

Over the last decade, tangential vector finite-element [1] and transfinite-element methods [2] have been developed for the solution of Maxwell equations. This discretization procedure results in a polynomial matrix equation in frequency. Especially in a waveguide structure, the modal fields of the ports depend strongly on frequency. To compute

Manuscript received October 11, 1999.
D.-K. Sun and Z. Cendes are with the Ansoft Corporation, Pittsburgh, PA 15219 USA.
J.-F. Lee is with the Electrical and Computer Engineering Department, Worcester Polytechnic Institute, Worcester, MA 01609 USA.
Publisher Item Identifier S 0018-9480(01)01087-0.