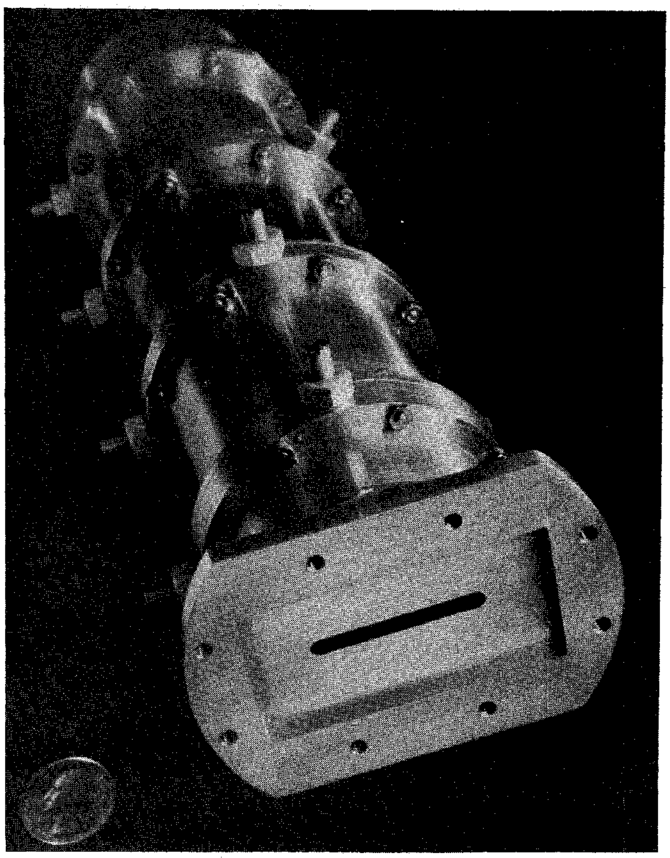

Fig. 9. Photograph of eight-pole filter.

Experimental results for two six- and eight-pole S-band elliptic function bandpass filters show good agreement with theory. Although this type of filter may suffer from spurious input–output coupling, it has been shown that careful design can alleviate this problem.

Acknowledgment

The authors are indebted to R. W. Kreutel for his constant encouragement during the development of this filter, and to R. Kessler, who constructed the experimental filters.

References

[1] S. B. Cohn, "Direct-coupled-resonator filters," *Proc. IRE*, vol. 45, pp. 187–196, Feb. 1957.

[2] J. D. Rhodes, "The generalized direct-coupled cavity linear phase filter," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-18, June 1970.

[3] A. E. Atia and A. E. Williams, "Narrow bandpass waveguide filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-20, pp. 258–265, Apr. 1972.

[4] ——, "Non-minimum phase optimum amplitude bandpass waveguide filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-22, pp. 425–432, Apr. 1974.

[5] A. E. Atia, A. E. Williams, and R. W. Newcomb, "Narrow-band multiple coupled cavity synthesis," *IEEE Trans. Circuits Syst.*, vol. CAS-21, pp. 649–655, Sept. 1974.

[6] R. Levy, "Filters with single transmission zeros at real or imaginary frequencies," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-24, pp. 172–181, Apr. 1976.

[7] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance Matching Networks and Coupling Structures.* New York: McGraw-Hill, ch. 5.

Components for Microwave Integrated Circuits with Evanescent-Mode Resonators

KLAUS SCHÜNEMANN, MEMBER, IEEE, REINHARD KNÖCHEL, AND GÜNTHER BEGEMANN

Abstract—**The electrical performance of active microwave components for radio link systems, which have been realized utilizing evanescent-mode resonators, is described. This waveguide-below-cutoff technique is shown to be an alternative to the techniques established before now.**

I. Introduction

A FEW YEARS AGO, Craven [1] presented a new type of passive integrated circuitry utilizing evanescent-mode resonators. In this technique, inductance is represented by short sections of rectangular waveguide below cutoff, capacitance by obstacles in the waveguide, such as a capacitive screw or a thin sheet of dielectric. Thus resonators of high unloaded Q-factor (called evanescent mode resonators) can be formed which resemble a reentrant cavity (see, e.g., [2]) in that the electric stored energy is confined to a small volume of a gap region surrounded by a larger volume, which contains the magnetic stored energy. In both cases resonant conditions are established only after insertion of a post. One can then regard the reentrant cavity as the forerunner of the waveguide-below-cutoff technique.

While theory and realization of passive components in waveguide-below-cutoff technique are in an advanced state [2]–[10], only little work has been done concerning active components [11]–[15]. A varactor diode upconverter and

Manuscript received May 11, 1977; revised August 10, 1977. This work was supported by the Deutsche Forschungsgemeinschaft.

The authors are with the Institut für Hochfrequenztechnik, Technische Universität Braunschweig, D-3300 Braunschweig, Federal Republic of Germany.

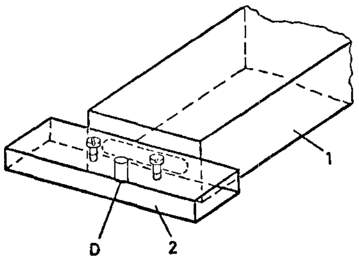

Fig. 1. Prototype oscillator circuit. D-active device, 1-propagating waveguide, 2-evanescent mode resonator.

several types of Gunn oscillators have been described, which show no outstanding electrical performance. In particular, the inherent advantages of the new technique for realizing active components have not been utilized or at least recognized. Hence, this work is devoted to demonstrating the applicability and superiority of the waveguide-below-cutoff technique for high-quality microwave systems such as microwave radio links and satellite systems.

II. Prototype Circuit

For application purposes, we have chosen radio link system components utilizing both FM and phase-shift keying (PSK). In order to achieve a high degree of standardization, a prototype oscillator circuit has been developed which can be used for local oscillators, injection locked amplifiers for both FM and PSK signals, voltage controlled oscillators for use in a phase-locked loop for amplification or frequency stabilization, and cavity-stabilized oscillators. The circuit is suitable for IMPATT diodes and Gunn elements. The prototype circuit can further be used as phase modulator with pin or varactor diodes for performing PSK.

The circuit is sketched in Fig. 1. The resonant diode mount is formed by the evanescent-mode waveguide 2, the power is extracted through a propagating waveguide 1. Both waveguides are coupled by a narrow slot in their common wall. (In an integrated-circuit system the propagating waveguide 1 will be replaced by the component subsequent to the oscillator, e.g., by a filter or by a circulator, which are realized by evanescent mode resonators, too.) Resonance conditions are established only after introduction of the active device D, which stores predominantly electric energy, and thereby establishes the necessary balance between time-average amounts of electric and magnetic stored energies. The resonance structure is thus limited to the volume immediately surrounding the active device. The screws can fulfill either of two operations: They can tune the oscillation frequency if they are located in the vicinity of the active device; otherwise they form additional resonators as is important, e.g., if one wants to increase the locking bandwidth of an injection locked amplifier for FM signals by reactance compensation. The load is matched via the dimensions of the slot, which but weakly affects frequency.

III. Basic Features

The significant features of the technique are the following: The individual elements of equivalent circuits approximate lumped elements in a large frequency range. There is no periodic relationship between reactance and frequency as in other distributed constant networks, for progressive waves are not supported if the frequency does not exceed the cutoff frequency of the host waveguide. In a frequency range where it does, however, poles and zeros of network functions will in general not be harmonic to the working frequency, and hence will not degrade the electrical performance. Broadband performance should therefore be characteristic for active components with evanescent mode resonators.

This important feature is a consequence of the lumped-element character of networks. It shall now be stressed once more, by regarding the problem of matching an impedance (e.g., of a semiconductor diode) over a large frequency range. If the transforming network is composed of lumped elements only, its bandwidth for a given mismatch is governed by the well-known relation of Fano [16]. The bandwidth around the working frequency degrades if lumped elements are replaced by distributed ones. This is mainly due to parasitic passbands, which are inherent in networks with distributed elements. Another undesired feature is that these passbands are harmonically related to the working frequency. The situation is quite different for networks with evanescent-mode resonators: As long as the frequency does not exceed the cutoff value of the host waveguide, additional passbands do not exist. At higher frequencies, parasitic passbands are indeed possible; they do not fall, however, at harmonics of the working frequency. Moreover, they are in general not so strongly marked as in the case of distributed constant networks. Hence, their contribution to Fano's mismatch integral is smaller, thus leading to a larger bandwidth of the fundamental passband. With respect to bandwidth, evanescent mode resonators yield medium performance between the performance of lumped-element and distributed-element networks.

Another advantage of the technique is that a given equivalent circuit can easily be realized without introducing undesirable parasitic elements as, e.g., in normal waveguide technique. This is likewise due to the lumped character of both the inductances and capacitances. Both elements concentrate their associated fields in a small and well-defined volume. Moreover, as almost all types of discontinuities inserted in a waveguide below cutoff behave like inductances, their effect can easily be included in the resultant inductance of the waveguide. Hence there are no problems with parasitic elements of capacitive character. Thus a high degree of predictability is achieved when special components are realized.

Further advantages of the waveguide-below-cutoff technique are a substantial reduction in size, weight, and cost of both passive and active microwave components. The technique does not, of course, compete with existing thin- and thick-film techniques in this respect; it leads, however, to superior electrical performance.

Concluding this chapter, the significant features of the waveguide-below-cutoff technique shall be summarized:

1) lumped element character of networks;
2) broad-band performance;

3) high degree of predictability;
4) lack of parasitic elements of capacitive character;
5) small circuit losses;
6) feasibility of integrating passive and active components;
7) remarkably simple microwave networks;
8) substantial reduction in size, weight, and cost.

IV. Applications

In the following the potentials of the waveguide-below-cutoff technique will be demonstrated by describing the performance of active microwave components. The experimental results presented below are average values, and typical for the special devices; they could sometimes have been surpassed considerably, if the circuit had been optimized for a single semiconductor element.

A. *Injection-Locked Amplifier*

Injection-locked solid-state oscillators can be used as amplifiers for FM signals replacing TWT's. System requirements then put severe restrictions on the amplitude and phase responses, the group delay variation, AM to PM conversion, and AM compression. Low FM distortion, i.e., a nearly linear relationship between phase ϕ and frequency ω at the amplifier output port, turns out to be the most important design objective.

The characteristic features of injection-locked oscillators are commonly illustrated in the admittance plane, showing curves of the frequency-dependent admittance of the passive circuitry (the load line), and of the RF-amplitude and frequency-dependent admittance of the active device (the device line). Stable oscillation is then represented by the intersection of device and load line, as is well known. Any injection signal can now be illustrated by an injection admittance, whose phase versus frequency relationship $\phi(\omega)$ can be graphically constructed from the admittance locus chart as shown by Kurokawa [17].

The inverse problem must be solved if an injection locked amplifier is to be theoretically optimized. Starting from a linear relationship between ϕ and ω, the corresponding load line must be found. In case of a van der Pol-type nonlinearity it turns out to be circular, with equal frequency spacing around the free-running oscillation point. The question of how to realize this passive circuitry arises now. An approach, which but insignificantly multiplies the number of tuning elements, represents a double-tuned oscillator, which consists, e.g., of a series resonant circuit in parallel to a parallel resonant circuit. The design parameters in this case are the load conductance and the coupling coefficients and Q-factors of the two circuits, which should be tuned to the same natural frequency. A still closer approximation to the ideal load line is obtained with a triple-tuned oscillator consisting, e.g., of three parallel resonant circuits in series, whose natural frequencies should all be different from one another. This circuit is believed to yield the best compromise between ideal behavior and need for circuit simplicity.

So far, only the shape of the load line has been fixed, whereas the load conductance at the oscillation frequency of the free-running oscillator must still be determined. A favorite choice is to establish the loading of the active device in such a way that maximum output power is delivered only in the state of injection locking at the free-running frequency. Compared to the case that the load is adjusted to maximize the output power of the free-running oscillator, both an improved locking bandwidth and less power variation with frequency is observed. (Of course the load has to be readjusted when the locking gain is changed.)

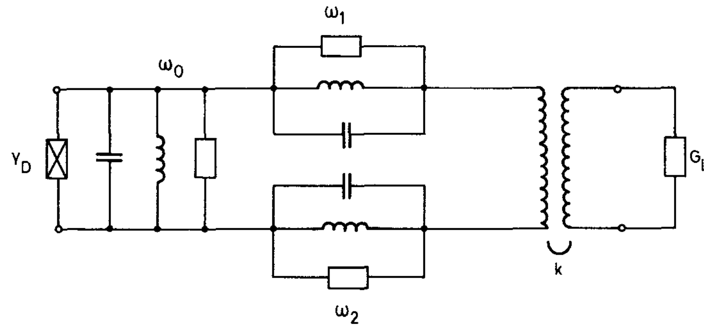

Fig. 2. Equivalent network of the prototype oscillator circuit. Y_D admittance of active device; G_D load conductance; k turns ratio of the coupling slot transformer; $\omega_0,\omega_1,\omega_2$–resonant frequencies ($\omega_1 < \omega_0 < \omega_2$ for a triple-tuned injection locked amplifier).

Following these theoretical guidelines, a triple-tuned oscillator has been developed with evanescent-mode resonators. It consists of the prototype oscillator circuit of Fig. 1. The three resonant circuits are each realized by a tuning screw (or the active device, respectively), and the surrounding sections of the waveguide below cutoff. The cutoff frequency has been chosen as 3/2 times the operating frequency; the waveguide height corresponds to that of the diode package. An equivalent network of the oscillator circuit is shown in Fig. 2. The various conductances of the parallel resonant circuits account for the circuit losses. The coupling of the circuits is adjusted by the distances between the tuning screws and the active device; the resonant frequencies are determined by the penetration depth of the screws. Adjustment of the load is done by the dimensions of the coupling slot.

Using a Gunn element, a relative locking bandwidth of 20 percent could be achieved at a gain of 10 dB with an output power variation of less than ± 0.5 dB. In case of 20-dB gain, the locking bandwidth exceeded 10 percent. A typical curve is given in Fig. 3. Phase linearity and AM to PM conversion were sufficient over a frequency range of 7 percent, i.e., group delay variation was less than 1 ns, and AM to PM conversion less than 4°/dB.

The locking bandwidth of an IMPATT diode oscillator was smaller due to the higher Q-factor of the IMPATT diode. At 20-dB gain, a 200-MHz locking bandwidth could be measured in the 7-GHz band. The amplifier could be tuned by one screw over a 10-percent frequency range with the locking bandwidth not less than 120 MHz (see Fig. 4).

The performance of injection-locked amplifiers with evanescent-mode resonators is largely affected by the harmonic content of the output spectrum. The aforementioned

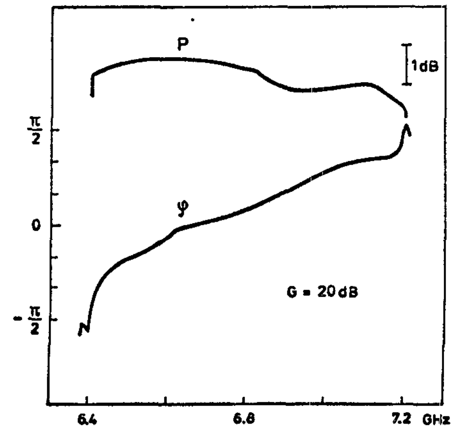

Fig. 3. Output power P and phase ϕ of injection locked Gunn oscillator versus frequency at 20-dB gain (maximum output power 22 dBm).

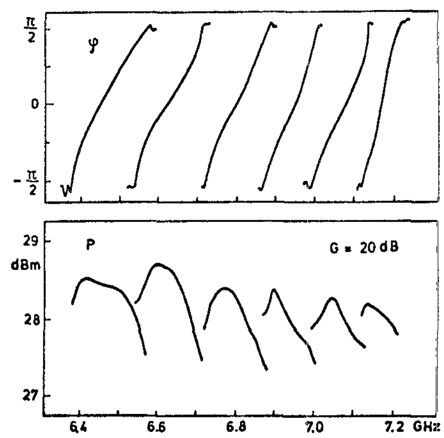

Fig. 4. Output power P and phase ϕ of injection-locked IMPATT oscillator versus frequency at 20-dB gain (the various curves have been obtained by tuning one screw).

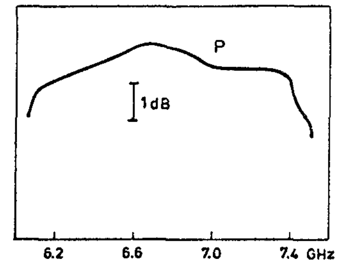

Fig. 5. Output power P of mechanically tuned Gunn oscillator (maximum output power 22 dBm).

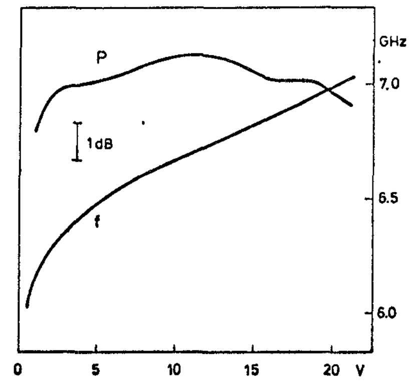

Fig. 6. Output power P and frequency f of electronically tuned Gunn oscillator versus varactor bias (maximum output power 21 dBm).

results could only be achieved provided that harmonics in the output spectrum had sufficiently been damped by applying harmonic absorbers to the ends of the evanescent-mode waveguide. If the harmonic absorbers are omitted, ripples will be found in the phase response of the injection locked oscillator. Keeping in mind that the slope of the phase response with frequency is a measure for the loaded Q-factor of the circuit, one can state that these ripples indicate a Q-factor degradation at some discrete frequencies. This degradation is caused by reactive power stored at harmonics of the oscillation frequency. The impedance locus seen from the active device then shows small loops. Although stable locking around such loops is possible provided that the locking power is not too small, ripples in the phase response and the involved variations in group delay limit the applicability of the oscillator as an FM amplifier. Employing harmonic absorbers is, however, a powerful means to overcome these difficulties.

B. Voltage-Controlled Oscillator

A voltage-controlled oscillator (VCO) has been realized with a Gunn element as active device. Locating the two screws of the prototype circuit in the vicinity of the active device yields an oscillator whose oscillation frequency can be tuned by one screw over 18-percent bandwidth with the output power being constant within ± 0.5 dB (see Fig. 5). Over a relative frequency range of 15 percent the output power ripple is even less than ± 0.25 dB.

Replacing one screw of the prototype circuit by a varactor diode yielded a tuning range of 15 percent with an output power ripple less than ± 0.5 dB. The power loss in the varactor diode was 0.5 to 1 dB. Typical tuning curves are given in Fig. 6. One problem occurs, however, in practice; when tuning the oscillator mechanically or electronically, instabilities are often observed over the tuning range which lead to frequency jumps of several 10 MHz. This effect heavily depends on the particular Gunn element used. It was found to be related to the harmonic content of the oscillator output spectrum. Because of the low Q-factor of the imbedding, even small amplitudes of the harmonics cause a storage of reactive power in the evanescent-mode waveguide. This leads to small impedance loops which can be observed by displaying the impedance locus seen from the active device. These loops are due to a further lowering of the Q-factor, which results in instabilities: Continuously tuning the VCO around a loop is not possible, a discontinuous frequency jump will occur instead. The problem has been solved by applying harmonic absorbers to the ends of the evanescent-mode waveguide. The performance of the VCO is then free from mode jumping effects.

A mechanically and electronically tunable oscillator showed the following characteristic features: The oscillation frequency could be mechanically tuned over a relative

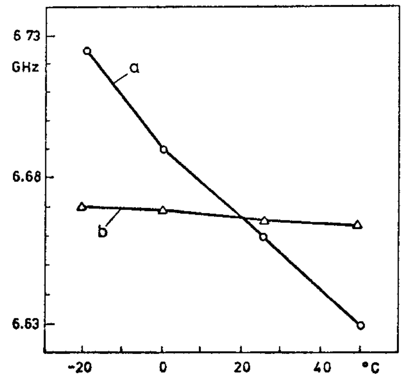

Fig. 7. Oscillation frequency versus temperature of a Gunn oscillator with varactor diode without (a) and with (b) a compensating dielectric.

bandwidth of 13 percent (output power constant within ±0.5 dB), and additionally by a varactor diode over 5 percent (additional power ripple less than ±0.25 dB). Inserting another varactor diode for FM purposes yielded tuning ranges of 15 percent (mechanically), and of 5 percent and 30 MHz, respectively, for the two varactor diodes. The 30-MHz tuning range allowed performing linear FM.

The temperature dependence of the oscillation frequency is shown in Fig. 7. The frequency drift can effectively be reduced by placing a small slice of dielectric with a negative temperature coefficient of its dielectric constant in the gap under the tuning screw. Thus a stabilization factor of 20 could be achieved.

C. Cavity-Stabilized Oscillator

Cavity stabilization is well known as an effective and simple means for improving both short-term and long-term frequency stability of solid-state oscillators. The compound oscillator structure consists of three networks, which are coupled together: the diode mounting structure or the original oscillator, the high-Q resonator, and the coupling network (transmission line with damping resistor). The passive circuitry must carefully be designed in order to prevent multiple-mode operation due to the interaction of the several resonant structures. To this end an appreciable amount of the generated power (typically 1–2 dB) must be absorbed in the damping resistor. A further disadvantage is that the oscillator can only be tuned over a narrow frequency range by varying the natural frequency of the cavity. This is due to the phase sensitivity with frequency of the coupling line. Until now, single-mode operation of the compound oscillator structure has been reported over a relative frequency range of maximally 1 percent for stabilization factors ranging between 10 and 100. To improve the performance of the cavity-stabilized oscillator assembly the governing design objective has to be the renunciation of the coupling line. This has not yet been achieved for oscillators in normal waveguide, coaxial, or MIC technology, but it can indeed be achieved by utilizing waveguides below cutoff.

Regarding our prototype circuit, the stabilizing cavity can be mounted adjacent to the evanescent-mode waveguide. A

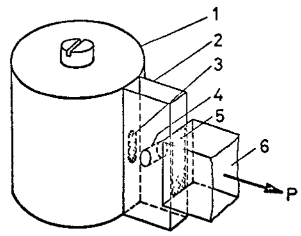

Fig. 8. Reflection cavity stabilized Gunn oscillator. 1–TE_{011}-mode circular waveguide cavity; 2–evanescent mode resonator; 3,5–coupling holes; 4–active device; 6–propagating waveguide.

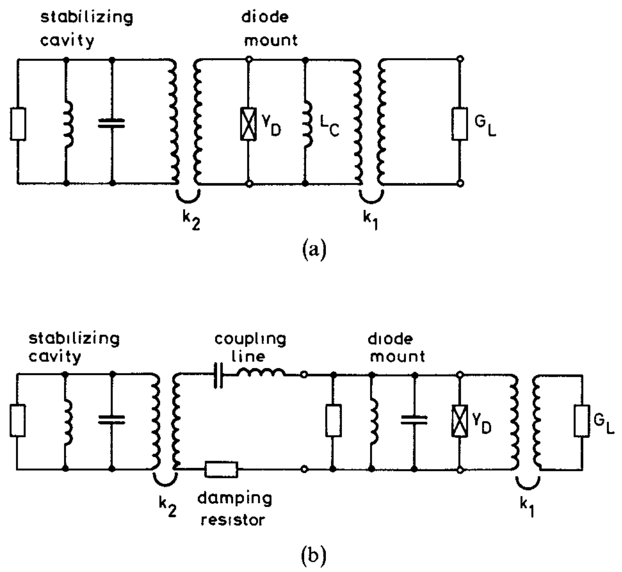

Fig. 9. Equivalent network of a cavity-stabilized oscillator. (a) With evanescent-mode resonator. (b) In normal waveguide technique. Y_D–admittance of active device; G_L–load conductance; L_c–resultant inductance of waveguide below cutoff; k_1,k_2–turns ratio.

possible configuration of a reflection-cavity stabilized oscillator is schematically sketched in Fig. 8. No further tuning screw is needed in this case, so that the evanescent-mode waveguide performs as a pure inductance as seen from the active device. It can hence be represented in an equivalent circuit (although somewhat idealized) as a transformer which connects the load and the stabilizing cavity to the active device (see Fig. 9(a)). The transformation ratios can be separately adjusted for load and cavity by choosing the dimensions of the coupling slots. The compound oscillator structure is described by a simple parallel resonant circuit. This holds in a frequency range of up to 10 percent, in which single-mode operation can be obtained. In the case of cavity-stabilized oscillators realized in conventional techniques, the equivalent circuit contains 3 resonant circuits for the diode mounting structure, the high-Q resonator, and the coupling line (see Fig. 9(b)). Single-mode operation is thus restricted to a far narrower frequency range.

For the reflection cavity-stabilized oscillator of Fig. 8 the power is extracted through a propagating waveguide. In a modification of this circuit, the power can be extracted through another port of the cavity. This leads then to a transmission cavity-stabilized oscillator.

Using a circular TE_{011}-mode waveguide cavity, a

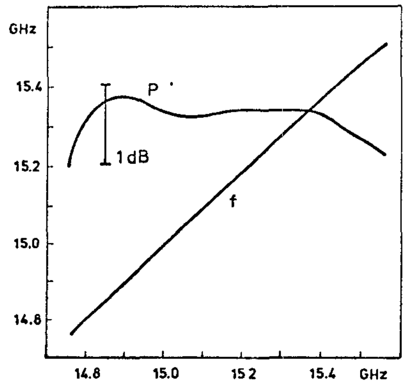

Fig. 10. Output power P and oscillation frequency f versus cavity frequency of a reflection cavity-stabilized Gunn oscillator (maximum output power 260 mW).

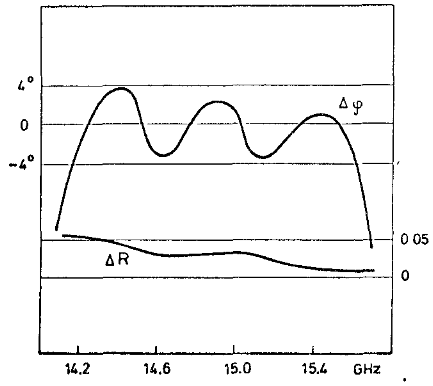

Fig. 11. Phase shift error $\Delta\phi$ and imbalance ΔR of the reflection coefficients of a phase modulator with pin-diode.

reflection-cavity-stabilized oscillator has been realized at *Ku*-band (15 GHz). The loaded Q-factor of the oscillator amounted to 4500 with a power loss in the cavity of less than 2 dB (unloaded Q-factor of the cavity being 20000 at 15 GHz). Over a tuning bandwidth of 700 MHz the output power was constant within ± 0.5 dB (see Fig. 10). The tuning range is only limited by the mechanical tunability of the cavity. An additional electronic tuning range of 10 MHz could be achieved by hole-coupling another evanescent-mode resonator with varactor diode to the high-Q cavity. This tuning circuit absorbed 1 dB of power and simultaneously reduced the loaded Q-factor to 3700. The FM noise at modulation frequencies exceeding 10 kHz was less than 0.1 Hz measured in a 100-Hz bandwidth. By mechanically compensating for the frequency drift with temperature change, a frequency stability of 1×10^{-5} could be achieved from -20 to $+50$°C. To this end the high-Q cavity was made from Invar, with its bottom from copper, and filled with dry nitrogen.

D. Phase Modulator with Pin-Diode

Phase modulators with pin-diode are advantageous for PSK in both radio-link systems and guided millimeter-wave communication systems with high bit rate. A severe restriction in the performance of these modulators is their narrow bandwidth of typically 1 percent, which is defined with respect to a tolerable deviation from the desired phase shift (e.g., $\pm 4°$ in case of an 180° modulator) and to a tolerable imbalance in the magnitudes of the reflection coefficients of the two diode states (less than, e.g., 0.05).

A reflection-type phase modulator has been developed by using the prototype circuit of Fig. 1 with a pin-diode as nonlinear device D. The imbedding is capable of performing a broad-band resonance transformation of the two pin-diode impedance states in such a way, that reflection coefficients of 180° phase difference and of equal magnitudes are achieved. The modulator bandwidth amounted to 10 percent for an insertion loss of 0.5 dB and 1-ns switching times at a midband frequency of 14.9 GHz. Phase-shift error $\Delta\phi$ and imbalance ΔR of the reflection coefficients are shown versus frequency in Fig. 11.

From the foregoing results it can be stated that the prototype circuit of Fig. 1 is well suited for performing a broad-band matching of various semiconductor elements. Gunn elements, IMPATT, varactor, and pin-diodes can be successfully incorporated.

V. Conclusions

The significant features of the waveguide-below-cutoff technique shall be summarized and compared to those of conventional techniques. Its circuit elements show lumped character. Hence broad-band performance is characteristic of realized active components. The technique is in this respect superior to coaxial, normal waveguide, and thin and thick film techniques. As parasitic elements of only inductive character are introduced in a realization (which can be absorbed in the inductance of the waveguide below cutoff), remarkably simple microwave networks can be established, which are further characterized by a high degree of predictability. This is an advantageous difference to normal waveguide techniques. Moreover, the waveguide-below-cutoff technique is superior to coaxial or to normal waveguide technique in the sense of cost, size, and weight of realized components. It is hence believed to represent a remarkable alternative to the techniques established till now. It should be applicable up to frequencies in the lower millimeter-wave region (30–40 GHz).

References

[1] G. F. Craven, "Waveguide below cutoff: A new type of microwave integrated circuit," *Microwave J.*, vol. 13, p. 51, Aug. 1970

[2] A. G. Williamson, "The resonant frequency and tuning characteristics of a narrow-gap reentrant cylindrical cavity," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-24, pp. 182–187, Apr. 1976.

[3] W Haas, "Open electromagnetic resonators," *Arch. Elek. Übertragung*, vol. 18, pp. 395–402, July 1964

[4] W. Haas and H. D. Godtmann, "Microwave bandfilters using one-dimensional open resonators," *Arch. Elek. Übertragung*, vol 19, pp. 551–558, Oct. 1965.

[5] G. F. Craven, "Waveguide bandpass filters using evanescent modes," *Electron Lett.*, vol. 2, no. 7, pp. 251–252, July 1966.

[6] G. F. Craven and C. K. Mok, "The design of evanescent mode waveguide bandpass filters for a prescribed insertion loss characteristic," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-19, pp. 295–308, Mar. 1971.

[7] C. K. Mok, D. W. Stopp, and G. F. Craven, "Susceptance-loaded evanescent mode waveguide filters," *Proc. Inst. Elec. Eng.*, vol. 119, pp. 416–420, Apr. 1972.

[8] C. K. Mok, "Design of evanescent mode waveguide diplexers," *IEEE*

Trans. Microwave Theory Tech., vol. MTT-21, pp. 43–48, Jan. 1973.
[9] R. F. Skedd and G. F. Craven, "New type of magnetically tunable multisection bandpass filter in ferrite-loaded evanescent waveguide," *IEEE Trans. Magn.*, vol. MAG-3, pp. 397–401, Sept. 1967.
[10] L. F. Franti, "4 GHz low noise parametric amplifier," in *Proc. 6th Eur. Microwave Conf.* (Rome, Italy), pp. 545–551, 1976.
[11] W. Kwiatkowski, T. Arthanayake, and V. H. Knight, "Efficient high level upconverter for radio link," *Electron. Lett.*, vol. 5, no. 19, pp. 625–627, Sept. 1970.
[12] F. Ivanek, M. Shyam, and V. G. K. Reddi, "Investigation of waveguide-below-cutoff resonators for solid-state active devices," *Electron. Lett.*, vol. 5, no. 10, pp. 214–216, May 1969.
[13] P. S. Kooi and D. Walsh, "Novel technique for improving the frequency stability of Gunn oscillators," *Electron. Lett.*, vol. 6, no. 4, pp. 85–86, Feb. 1970.
[14] P. G. Wilson and B. Minakovic, "Development of an FM pulsed Gunn oscillator at X-band," *IEEE Trans. Electron Devices*, vol. ED-18, p. 450, July 1971.
[15] R. F. B. Conlon, R. E. Cooke, and J. S. Heeks, "Broadband injection locking of Gunn oscillators," in *Proc. 3rd Eur. Microwave Conf.* (Brussels, Belgium), p. A.3.1, 1973.
[16] R. M. Fano, "Theoretical limitations of the broadband matching of arbitrary impedances," *J. Franklin Inst.*, vol. 249, pp. 57–84 and 139–154, Jan.–Feb. 1950.
[17] K. Kurokawa, "Injection locking of microwave solid-state oscillators," *Proc. IEEE*, vol. 61, pp. 1386–1410, Oct. 1973.

Short-Circuit Tuning Method for Singly Terminated Filters

MING HUI CHEN, MEMBER, IEEE

Abstract—**This paper describes a filter tuning method based upon the match of measured and computed input impedances for a short-circuited filter. Two singly terminated filters, an 8-pole Chebyshev filter, and a 6-pole pseudoelliptic function filter tuned by using this method have demonstrated excellent performance.**

Introduction

Singly terminated filters are the key elements in constructing a contiguous band multiplexer [1]. Due to the lack of a systematic tuning method for the singly terminated filter, the contiguous band multiplexer would be difficult to construct and tune. This may be why the contiguous band multiplexer is seldom used in a practical system.

The conventional filter tuning is based upon the return loss characteristics of a filter. Minimum reflection in the passband with correct center frequency and bandwidth is usually the criterion for tuning doubly terminated filters. Since the input port for a singly terminated filter is not matched over the entire passband due to the existence of passband reactance, the criterion of minimum reflection cannot be used for filter tuning. A method based upon the filter's short-circuit impedance, which has been used for the measurement of intercavity couplings [2], has been further developed for tuning singly terminated filters. This paper presents this filter tuning method, as well as the experimental results for two model filters tuned by using this method.

Manuscript received May 16, 1977; revised July 18, 1977. This paper is based upon work performed in COMSAT Laboratories under the sponsorship of the International Telecommunications Satellite Organization (INTELSAT). Views expressed in this paper are not necessarily those of INTELSAT.

The author was with COMSAT Laboratories, Clarksburg, MD 20734. He is now with the Antenna and Communication Laboratory, TRW Systems Group, Redondo Beach, CA 90278.

Network Model for the Filter

With the short-circuit filter tuning method, the tuning and coupling screws are set one by one according to the match of measured and computed input impedances for a short-circuited filter. Therefore, the correct network representation for the filter is extremely important. Consider a multiple-coupled cavity network [3], [4], whose currents and voltages are related by an impedance matrix Z as follows:

$$\begin{bmatrix} V_1 \\ V_2 \\ \vdots \\ V_N \end{bmatrix} = Z \begin{bmatrix} I_1 \\ I_2 \\ \vdots \\ I_N \end{bmatrix} \tag{1}$$

with

$$Z_{ii} = \frac{1}{Q_u \times BW} + j\frac{2}{\pi \times BW}\left(\frac{\lambda_{0i}}{\lambda g_{0i}}\right)^2 \cdot \tan\left(\frac{\pi \lambda g_{0i}}{\lambda_g}\right), \qquad i = 1, 2, \cdots, N \tag{2}$$

$$Z_{ij} = jM_{ij}, \qquad i \neq j \tag{3}$$

where Q_u is the unloaded cavity Q, BW is the fractional bandwidth, and λ and λ_g are the free space and guide wavelengths, respectively.

When the cavity i is resonant, λ_{0i} and λg_{0i} are set according to the midband frequency of the filter; when i is not resonant, λg_{0i} and λ_{0i} are set according to a detuned frequency which may be measured in advance. The coupling coefficients, M_{ij}, the input resistance, R_1, and the output resistance, R_2, define the network as shown in Fig. 1 with the input at AA' and output at BB'. A short-circuited filter is a 1-port network with a short circuit at BB' and the input at