Embedding circuits were designed that provided the optimum collector and emitter terminations to the device. The optimum base termination is very close to a short circuit and, for convenience, the base terminal was shorted directly to ground. The circuits were fabricated on duroid using microstrip transmission lines and chip capacitors. A circuit diagram of the complete oscillator, including the bias network, is shown in Fig. 5. The measured output power of the oscillator was 17.5 dBm at a frequency of 2.69 GHz. This output power corresponds to a power-added efficiency of 32 percent.

IV. Conclusions

An empirical design technique for series-type oscillators has been described that maximizes output power. Using this technique, the characteristics of the embedding circuits required for maximum output power are measured directly under large-signal conditions similar to those encountered during actual oscillation. These measurements are made possible by the use of a "two-signal" technique which prevents the active device from oscillating at unwanted frequencies.

A bipolar transistor oscillator was constructed to verify the design technique. The predicted oscillation frequency of 2.7 GHz was very close to the actual frequency of 2.69 GHz. The output power of 17.5 dBm was 0.4 dB lower than the power predicted by the test system with the passive tuner and 1.5 dB lower than the power predicted by the test system with the sliding short. This is reasonable agreement; it is assumed that the loss in the output circuit of the completed oscillator was similar to the uncorrected loss in the power measurement with the single power meter and passive tuner. The output power was also close to the maximum added power of the active device when used as a common emitter amplifier. These results suggest that the large-signal conditions seen by the device during actual oscillation were established during the measurement and that the output power of the oscillator was close to the maximum power that could be achieved with the active device.

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Letters

Comments on "The Measurement of Noise in Microwave Transmitters"

WOLFGANG FREUDE

The recent comment by Ashley *et al.* [1] as a reply to the remarks by Knöchel *et al.* [2] drew my attention to the work by Ashley, Barley, and Rast, Jr. [3], to which I would like to add three supplements. On an improved version of the frequency discriminator of Ondria [4] and Ashley *et al.* [5], the following results have been published [6].

1) As can be seen by a linearized quasistatic analysis of the setup, there is no need for a matched pair of diodes. The adjustment for optimum AM noise suppression can be accomplished by a simple LF resistive network.

2) An AM noise suppression factor was defined which is independent of irrelevant discriminator parameters.

3) The conditions for the validity of the quasistatic approximation have been stated.

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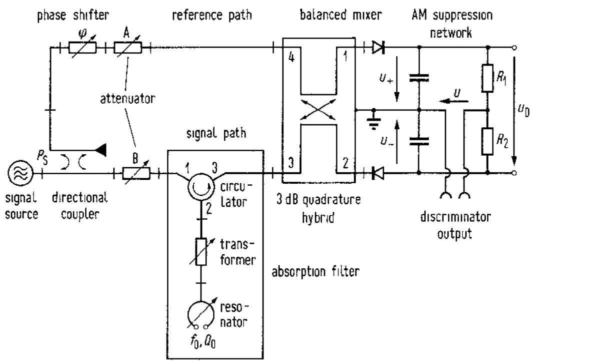

Fig. 1. Microwave frequency discriminator with AM suppression network R_1, R_2. u, discriminator output voltage; u_D, diode voltage.

To give these findings in more detail, the improved frequency discriminator of Fig. 1 has to be analyzed, yielding for the linearized quasistatic discriminator output voltage $u = u_R$ using a resonator at port 2 of the circulator [6]

$$u_R = \sqrt{P_S Z_L}\,\frac{\alpha_1 R_{S2} + \alpha_2 R_{S1}}{R_{S1} + R_{S2}}(-1)^k \sqrt{B}\,\frac{Q_0}{f_0}\Delta f + \sqrt{P_S Z_L}\,\frac{\alpha_1 R_{S2} - \alpha_2 R_{S1}}{R_{S1} + R_{S2}}\sqrt{A} \quad \text{(1a)}$$

If port 2 of the circulator in Fig. 1 is connected to a shorted transmission line, the output voltage $u = u_T$ becomes [6]

$$u_T = \sqrt{P_S Z_L}\,\frac{\alpha_1 R_{S2} + \alpha_2 R_{S1}}{R_{S1} + R_{S2}}(-1)^k \sqrt{\frac{AB}{A+B}}\,\frac{k\pi}{f_0}\Delta f + \sqrt{P_S Z_L}\,\frac{\alpha_1 R_{S2} - \alpha_2 R_{S1}}{R_{S1} + R_{S2}}\sqrt{A+B}. \quad \text{(1b)}$$

α_1, α_2 stand for the voltage rectification efficiencies of the diodes including all unsymmetries of the balanced mixer, $R_{S1} = R_1 + W_1$, $R_{S2} = R_2 + W_2$ result from the addition of the resistances R_1, R_2 in Fig. 1 to the video resistances W_1, W_2 of the diodes, P_S is the system input power, Z_L the line impedance, $A, B \leqslant 1$ are the power gains of attenuators A, B including all losses in the signal and reference arm, $k\pi$ is the phase shift of both arms at $\Delta f = 0$ including the variable phase shifter φ, $k = 0, \pm 1, \pm 2, \ldots, Q_0$ is the unloaded quality factor of the resonator, and $\Delta f = f - f_0$ is the frequency offset from the center frequency f_0. The dc level of the diode voltages u_+, u_- is assumed to be large, so the probability of overmodulation by the signal modulation remains low. It can be shown [6] that $u = u_R$, $u = u_T$ are functions of the ratio of the diode voltages u_+, u_- measured in an open-circuit condition, therefore Fig. 1 resembles a ratio detector. However, because the modulation converter is decoupled from the rectification diodes by an impedance matching network, the AM noise suppression is not independent from Δf, as it is approximately the case with the usual FM ratio discriminators. In that respect, the circuit of Fig. 1 has the properties of an ordinary push-pull slope detector, which suppresses AM noise only for $\Delta f = 0$, and if the second term in (1) disappears by choosing $\alpha_1 R_{S2} = \alpha_2 R_{S1}$.

The improvement of the present circuit [6] in Fig. 1 over the usual equipment [1]–[5] lies in the fact that the diodes need not be matched, neither in the voltage rectification efficiencies α_1, α_2 nor in the video impedances W_1, W_2. The balancing condition can be achieved very easily by adjusting the two LF resistances R_1, R_2, while Ondria [4] states that "there is no substitute for a well matched pair of diodes".

Fig. 2 shows the demodulation characteristic $u(f)$ of the adjusted circuit Fig. 1 with the parameters $A = -12.3$ dB, $B = -14$ dB (far from optimum to guard the diodes), $f_0 = 7$ GHz, $Q_0 = 7000$, $f_0/Q_0 = 1$ MHz, $P_S = 8.5$ mW, and nonmatched point-contact diodes 1N415G. The steep slope near f_0 is described by (1a), while the slope of the sinusoidal output voltage dependence (as to be seen in a transmission-line configuration or for $\Delta f \gg f_0/Q_0$ with fully reflecting resonator) is given by (1b). Near f_0, the demodulation sensitivity amounts to 0.244 mV/kHz with a near-linear range of ± 300 kHz. The lower trace in Fig. 2 shows that the sum u_D of the diode voltages remains constant, if the linear approximations for u as given in (1) are valid.

In relation to an equivalent simple slope detector of equal sensitivity (but different threshold), an AM voltage suppression factor S could be defined [6] independent of A

$$S = \frac{f_0}{Q_0} \cdot \frac{1}{|\Delta f|} + 1 \quad (2)$$

with the assumption $\alpha_1 = \alpha_2$, $R_{S1} = R_{S2}$ for the sake of simplicity. Using the data above and taking a static offset $\Delta f = 30$ kHz caused by drift, a voltage suppression factor $S = 31$ dB results.

Finally, the validity of the quasistatic approximation has been investigated [6] in terms of the maximum frequency component f_{LF} of the FM-noise modulated signal, where the maximum frequency deviation is f_H. For a sinusoidal time varying frequency offset $\Delta f(t) = f_H \cos 2\pi f_{LF} t$, any distortion of the demodulated discriminator output voltage disappears, if the conditions

$$f_H f_{LF} \ll (f_0/Q_0)^2, \qquad f_H \ll f_0/Q_0 \quad (3)$$

are fulfilled; f_0/Q_0 is the unloaded bandwidth of the resonator. The second inequality guarantees the static linearity. With the given data, the spectrum of linear detection extends up to $f_{LF} \ll 3.3$ MHz, if $f_H = 300$ kHz.

Using such a set-up, the correlation of AM and FM Gunn oscillator noise, and the non-gaussian noise near the oscillation threshold have been measured successfully [7].

Reply[1] *by J. Robert Ashley*[2]

These comments refer mostly to the Ondria [4] and Ashley *et al.* [5] papers in the 1968 Noise Special Issue of these TRANSACTIONS. As an historical comment, we learned of Dr. Ondria's work and he learned of our work by reading the completed journal in 1968. The analysis by Dr. Freude does use more modern analytic techniques and gives many details omitted to keep both of the 1968 papers to a reasonable length.

In 1968, we used a balanced mixer consisting of a magic tee and two detectors because the same hardware was and still is needed to do the AM noise measurement. In the evolution of our laboratory techniques, we have gone to separate setups for AM and FM noise. The FM noise measurement is based on using catalog balanced mixers. We rely on the manufacturers to match the diodes because this is needed to optimize several mixer performance characteristics. Certainly, in using readily available balanced mixers for FM noise measurements from 30 MHz to 95 GHz, we have not observed any difficulties caused by unbalanced diodes.

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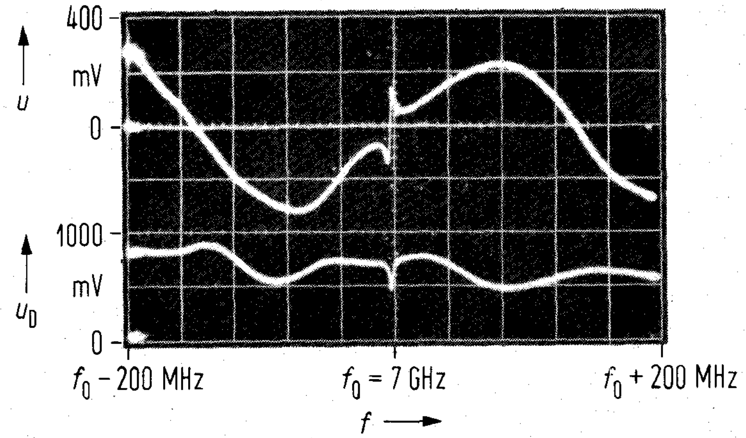

Fig. 2. Discriminator voltage u and diode voltage u_D as a function of frequency.

With respect to Dr. Freude's Figs. 1 and 2, the operation of his attenuator B at greater than zero dB is not desirable for making FM noise measurements. Since the absorption filter prevents the carrier power from overdriving the balanced mixer, attenuator B should have minimum loss to allow the noise sidebands to be as large as possible in order to over-ride the noise in the mixer diodes.

In [3], we called attention to the work of Fikart *et al.* [8], [9] on the effect of resonator detuning in the AM noise measurement. The analysis of noise measurement discriminators has been further advanced by Brozovich [10]. A new result on the bandwidth of the simplest of transmission-line discriminators was given at the 1983 MTT International Symposium by Ashley *et al.* [11].

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Comments on "The Design Parameters of Nonsymmetrical Coupled Microstrips"

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A recent paper [1] dealt with the subject of the design parameters of nonsymmetrical coupled microstrips for applications as microwave circuit elements. In this brief correspondence, the completeness and consistency of some of the results presented in the paper are examined and the existence of an earlier relatively comprehensive work reported on the same subject is pointed out [2].

First of all, it must be noted that, in general, the nonsymmetrical uniformly coupled inhomogeneous line system has six degrees of freedom, i.e., six independent variables are required to analyze and formulate design procedures for any circuit consisting of nonsymmetrical coupled microstrips [3], [4]. These six variables can be the equivalent line constants, that is, the self- and mutual-capacitances and inductances of the structure per unit length, or the six independent normal mode parameters derived from the line constants. There are eight normal mode parameters defined in [3], of which six are independent, e.g., β_c, β_π, R_c, R_π, Z_{c1}, and $Z_{\pi 1}$ or β_c, β_π, R_c, or R_π, and three of the four impedances. The other two variables can be determined from $Z_{c2}/Z_{c1} = Z_{\pi 2}/Z_{\pi 1} = -R_c R_n$ [3]. It should be mentioned that for certain special so-called congruent cases of coupled microstrips where even-voltage and odd-current modes can be defined, i.e., R_c is nearly equal to 1, only five variables need to be specified which can be the phase constants and three out of four mode impedances [3]. In the article under discussion, [1, fig. 7] representing the design data for a special but useful case of substrate material only gives impedances or three independent parameters. At least one of the mode–voltage ratios and both phase constants must also be specified for any analysis and design of a coupled-line circuit. The normal mode effective dielectric constants are given in [1, fig. 5] only for a special case of $S = 0.4$ mm in connection with the dispersion model.

From the statements made in the article [1] regarding the absence of any design data, it seems that the authors were not aware of an earlier study reported on the same subject [2]. In that

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