Wright-Patterson Air Force Base, Ohio, Tech. Rep. AFAL-TR-73-109, Mar. 1973.
[9] W. Baechtold, "*X*- and *Ku*-band amplifiers with GaAs Schottky-barrier field-effect transistors," *IEEE J. Solid-State Circuits* (*Special Issue on Microwave Integrated Circuits*), vol. SC-8, pp. 54–58, Feb. 1973.
[10] C. A. Liechti and R. L. Tillman, "Application of GaAs Schottky-barrier FETs in microwave amplifiers," in *ISSCC Dig. Tech. Papers*, 1973, pp. 74–75.
[11] H. Fukui, "Available power gain, noise figure, and noise measure of two-ports and their graphical representations," *IEEE Trans. Circuit Theory*, vol. CT-3, pp. 137–142, June 1966.
[12] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance-Matching Networks, and Coupling Structures.* New York: McGraw-Hill, 1964.
[13] A. F. Behle and R. Zuleeg, "Fast neutron tolerance of GaAs JFET's operating in the hot electron range," *IEEE Trans. Electron Devices* (Corresp.), vol. ED-19, pp. 993–995, Aug. 1972.

AM and FM Noise of BARITT Oscillators

JOSEF L. FIKART, MEMBER, IEEE

***Abstract*—AM, FM, and baseband noise of a BARITT diode oscillator in the range 100 Hz–50 kHz off the carrier has been measured under various operating conditions. A simple calculation has been made, relating the baseband noise to the oscillator AM and FM noise via measured amplitude and frequency modulation sensitivities and the results have been compared with the noise measured. It is shown that, depending on the bias current applied, both AM and FM noise performance can be degraded by up-conversion. Complete removal of up-converted noise requires a high-impedance low-noise bias supply since both the diode noise and bias supply noise at baseband frequencies may be significant when up-converted. Even with all modulation suppressed, the AM and FM noise has a flicker component almost completely correlated with the diode flicker noise at baseband frequencies. The RF power dependence of the AM and FM noise has also been investigated. It is shown that the BARITT oscillator noise compares very favorably with that of IMPATT's and TEO's. Values of −142 dB/100 Hz (AM noise) and 3.5 Hz/(100 Hz)$^{1/2}$ for $Q_{\text{ext}} = 200$ (FM noise) have been measured at 30 kHz off the carrier.**

I. INTRODUCTION

WITH their capability of producing low-noise microwave power in moderate amounts, BARITT diodes have become potential contenders to TEO's in local oscillator applications. Several theoretical studies of small-signal noise in BARITT's were carried out [1]–[3] and the principal noise sources (except for flicker noise) seem to have been identified. The results were given in terms of the small-signal noise measure. No theoretical large-signal noise analysis has been presented so far; perhaps the theories in [4],[5] or [6],[7], originally conceived for IMPATT's, could be used for this purpose once a suitable large-signal deterministic analysis is available for the BARITT diode. The experimental data on noise measure [8]–[11] indicate relatively small differences (10 dB on the average) between the small-signal and large-signal noise measures of BARRITT diodes.

However, there still is a considerable lack of information on the behavior of BARITT oscillator AM and FM noise, rather than noise measure, under various operating conditions. One reason for investigating AM and FM noise directly is the fact that the large-signal noise measure transforms differently into them [5],[6]; one can introduce the concept of AM and FM noise measure [5]. Another factor is the modulation noise which can sometimes completely mask the so-called "primary" noise, i.e., the internal noise generated by the device in the band about the center frequency. An effort in this direction was made by Herbst and Harth [11],[12] who measured the frequency modulation sensitivity of a Pd-n-p$^+$ BARITT diode in dependence on modulation frequency and bias current, in conjunction with FM noise measurements.

It is the purpose of this work to find out, by means of simple calculation and experiment, how both the total AM and FM noise are influenced by modulation noise, bias impedance levels, RF power level, etc., and to determine what minimum AM and FM noise can be expected and under what conditions. We will also be interested in seeing how the BARITT noise compares with that of other solid-state microwave generators.

II. CONDENSED THEORY

From the simple equivalent oscillator circuit in Fig. 1, and using well-known formulas for the primary noise deduced from [13],[14], the following relations can be written for AM and FM noise, respectively, while neglecting cross correlation between "primary" and modu-

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The author was with the Department of Electrical Engineering, the University of Alberta, Edmonton, Alta., Canada. He is now with GTE Lenkurt (Canada) Ltd., Burnaby, B. C., Canada.

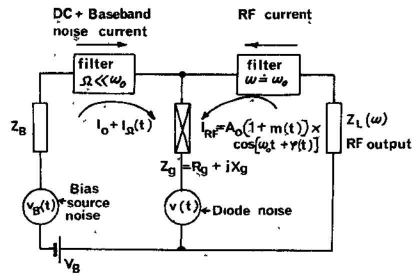

Fig. 1. Oscillator equivalent circuit.

lation noise:

$$(\mathrm{N/C})_{\mathrm{DSB}}{}^{\mathrm{AM}} = \underbrace{\frac{kTBM_{\mathrm{AM}}}{P_0(s/2)^2}}_{\text{primary}} + \underbrace{BS_{\mathrm{AM}}{}^2[\langle v_B{}^2(t)\rangle_\Omega + \langle v^2(t)\rangle_\Omega]/|Z_B + Z_g(\Omega)|^2}_{\text{modulation}} \quad (1)$$

$$(\Delta f_{\mathrm{rms}})^2 = \underbrace{\frac{f_0{}^2}{Q_{\mathrm{ext}}{}^2}\frac{kTBM_{\mathrm{FM}}}{P_0}}_{\text{primary}} + \underbrace{BS_{\mathrm{FM}}{}^2[\langle v_B{}^2(t)\rangle_\Omega + \langle v^2(t)\rangle_\Omega]/|Z_B + Z_g(\Omega)|^2}_{\text{modulation}} \quad (2)$$

where $(\mathrm{N/C})_{\mathrm{DSB}}{}^{\mathrm{AM}}$ is the double sideband noise-to-carrier ratio, Δf_{rms} is the rms frequency deviation, f_0 is the oscillation frequency, P_0 is the RF output power, Q_{ext} is the external quality factor, k is the Boltzmann's constant, $T = 290$ K, B is the noise bandwidth, M_{AM} and M_{FM} are the AM and FM noise measure, respectively, S_{AM} and S_{FM} are the AM and FM current modulation sensitivity, respectively, $\langle\ \rangle_\Omega$ denotes spectral density at a baseband frequency Ω, and the rest is obvious from Fig. 1, except for the nonlinearity factor s defined by [14]

$$s = \frac{A_0}{R_g}\frac{\partial R_g}{\partial A_0} \approx 2 \text{ (at full power)} \quad (3)$$

where A_0 is the magnitude of the RF current at the oscillation frequency f_0.

The noise measures M_{AM} and M_{FM} differ from the ordinary noise measure M because of the possible correlation of the sidebands of $v(t)$ under large-signal conditions [5],[6]. For the simple case of a phase stabilized oscillator and for low modulation frequencies

$$M_{\mathrm{AM}} = M(1 + |\rho_v| \cos x) \quad (4)$$

and

$$M_{\mathrm{FM}} = M(1 - |\rho_v| \cos x) \quad (5)$$

where

$$\rho_v = |\rho_v| e^{jx} \quad (6)$$

is the complex sideband correlation coefficient [5]. It is seen that (4) and (5) give the limits

$$M \le M_{\mathrm{AM}} \le 2M, \qquad 0 \le M_{\mathrm{FM}} \le M$$

or

$$0 \le M_{\mathrm{AM}} \le M, \qquad M \le M_{\mathrm{FM}} \le 2M. \quad (7)$$

Both $|\rho_v|$ and M, or more directly, the spectrum of the noise voltage $v(t)$ which determines M, generally increase with oscillator power level and this can be determined by measurement or calculation [5],[6] which, however, is somewhat involved and requires a computer. Here we shall only point out that the noise measure M of a nonlinear active device depends not only on its intrinsic noise around the oscillation frequency but also on that in the baseband (even if up-conversion through modulation is eliminated) and on the intrinsic noise at the oscillation frequency harmonics:

$$M = \sum_{l=-\infty}^{\infty} C_{1,l} S_{FF}(l\omega_0 + \Omega), \qquad \Omega \ll \omega_0 \quad (8)$$

where $S_{FF}(\ )$ is the power spectral density of the intrinsic noise process and $C_{1,l}$ are coefficients depending on the oscillator power level and various active device and circuit parameters. The cross-correlation factor can be also expressed by means of these coefficients [5]. It is seen that the term "primary noise," as applied to the first terms in (1) and (2), is somewhat inappropriate because of the interconversions indicated in (8).

The large-signal theories of [5] or [6] were applied to IMPATT diodes and good agreement was obtained with measurements. There are no theoretical data for large-signal BARITT noise as yet, but measurements indicate that the large-signal noise measure M is approximately in the range 21–29 dB as against the small-signal M of approximately 12–20 dB. This is a relatively small change compared to IMPATT's where M can typically increase from the small-signal value of some 34 dB to some 60 dB or even more.

III. CALCULATION OF THE PRIMARY AND MODULATION NOISE

We shall now estimate the individual contributions to the AM and FM noise for the BARITT diode used in the measurements described in Section IV. The diode parameters are given in Table I. For this diode, the AM and FM current modulation sensitivity was measured in the bias current range 10–20 mA (Fig. 2). Incidentally, these data cannot be directly compared to those in [12], where unfortunately, no reference to the external Q was made and also, no adjustments for maximum output power were applied at bias current values lower than the maximum current used. The curve of FM sensitivity versus bias current looks rather different if an effort is made (as is the case with the present data) to retune the circuit for maximum power output at each value of bias current. The curve of RF power versus bias current, measured under said conditions, is given in Fig. 3.

TABLE I
DIODE PARAMETERS

p+-n-p+ BARITT diode	
Area	$3\times10^{-4}cm^2$
n-region doping	$1.2\times10^{15}cm^{-3}$
Width	~ 8μ
Reach-through voltage	~ 56V
Max. dc current	50 mA
Max. output power (at 6.5 GHz)	20 mW

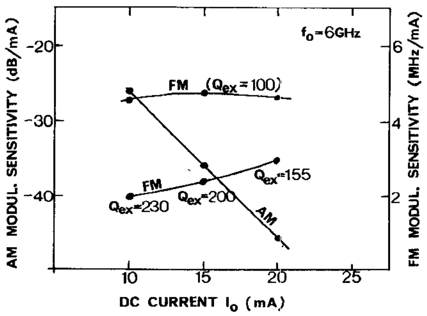

Fig. 2. AM and FM current modulation sensitivities as functions of bias current. (RF output power adjusted to maximum for each value of bias current.)

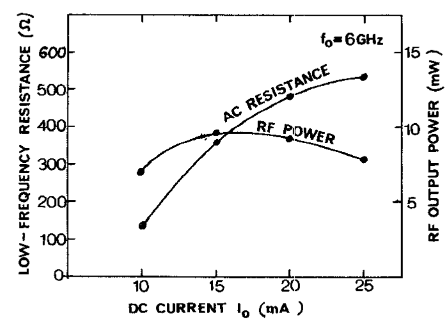

Fig. 3. RF power and low-frequency ac resistance (measured at 30 kHz) as functions of bias current. (RF output power adjusted to maximum for each value of bias current.)

The low-frequency ac resistance of the diode was also measured (Fig. 3) as it will be needed in (1) and (2). Finally, a measurement was made of the spectral density of the diode noise voltage, both due to the internal noise mechanism in the diode and due to the noise from the bias supply. The latter contribution was found to be relatively high even if a low-noise current supply was used (Fig. 4). To be able to measure the diode noise proper, an additional filter had to be used with the current supply as indicated in the insert of Fig. 4. As expected, a fair amount of flicker noise is seen to be present in the pure diode noise.

Let us now calculate the modulation noise for $I_0 =$ 15 mA, where maximum output power was obtained with this particular diode, and at modulation frequency 30 kHz. Assuming ideal noiseless bias supply, the AM noise due to the up-conversion of the diode low-frequency noise, for $Z_B = 0$, will be (in decibels)

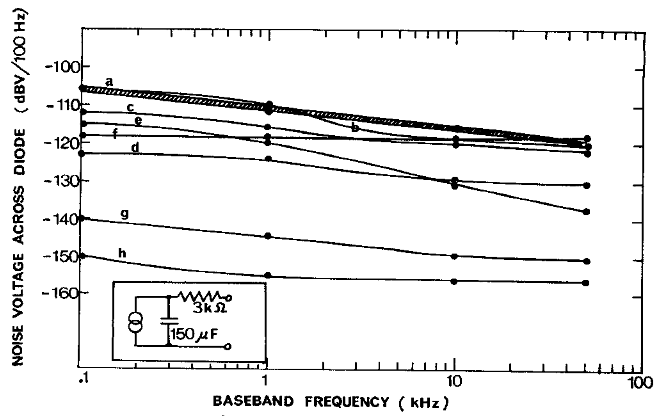

Fig. 4. Low-frequency diode noise voltage versus frequency. Curve a shows open circuit diode noise voltage for I_0 = 10–20 mA. Curves b–h show the noise voltage across the diode ac resistance due to the following:

curve b: current supply at I_0 = 20 mA, no filter;
curve c: current supply at I_0 = 15 mA, no filter;
curve d: current supply at I_0 = 10 mA, no filter;
curve e: voltage supply at I_0 = 15 mA, no filter;
curve f: voltage supply at I_0 = 10 mA, no filter;
curve g: current supply at I_0 = 10 mA, with filter (see insert);
curve h: current supply at I_0 = 15 mA, with filter (see insert).

$$\begin{aligned}(\mathrm{N/C})_{\mathrm{DSB}}{}^{\mathrm{AM(mod)}} &= 10\log\left[S_{\mathrm{AM}}{}^2 B\,\frac{\langle v^2(t)\rangle_\Omega}{R_{\mathrm{ac}}{}^2}\right]\\ &= 20\log S_{\mathrm{AM}} + 10\log B\langle v^2(t)\rangle_\Omega\\ &\quad - 20\log R_{\mathrm{ac}} = (-36+60)\\ &\quad - 118 - 51 = -145\ \mathrm{dB/100\ Hz}.\end{aligned}$$

(The 60-dB term arises from the need to express S_{AM} in decibels per ampere.) This can be compared to the primary noise expected. Taking $M_{\mathrm{AM}} = 20$ dB, $s = 2$, $P_0 = 10$ mW, $T = 290$ K, $B = 100$ Hz, and $k = 1.38 \times 10^{-23}$ J/K, we obtain

$$(\mathrm{N/C})_{\mathrm{DSB}}{}^{\mathrm{AM(prim)}} = 10\log\frac{kTBM_{\mathrm{AM}}}{P_0(s/2)^2} = -144\ \mathrm{dB}.$$

Hence, if $M_{\mathrm{AM}} = 20$ dB is close to reality, we can expect a 3-dB increase in AM noise as the primary and modulation noise are almost equal. Also, as can be seen from Fig. 4, the bias supply noise is lower than that of the diode so that the total noise increase should remain close to the above 3 dB. This is not very much. For $I_0 = 20$ mA, the modulation noise should be completely negligible since the modulation sensitivity is by some 10 dB lower than that for $I_0 = 15$ mA. However, the situation is entirely different for $I_0 = 10$ mA in which case the AM modulation sensitivity is by some 10 dB higher than at 15 mA and R_{ac} is smaller. Here the modulation noise will be more than 10 dB higher than the primary noise and the contribution due to bias supply noise will not be negligible.

Assuming again ideal noiseless bias supply with zero output impedance (ideal voltage source), the FM noise due to up-conversion of the diode low-frequency noise will be (I_0 = 15 mA, frequency 30 kHz)

$$(\Delta f_{\mathrm{rms}})^{\mathrm{mod}} = \frac{S_{\mathrm{FM}}}{R_{\mathrm{ac}}} (B\langle v^2(t)\rangle_\Omega)^{1/2} = \frac{1.2 \times 10^9}{3.6 \times 10^2} \cdot \left[\mathrm{antilog}\left(-\frac{118}{20}\right)\right]^{1/2} = 16.8\ \mathrm{Hz}$$

for a 100-Hz bandwidth and $Q_{\mathrm{ext}} = 100$.

The primary noise, taking $M_{\mathrm{FM}} = 25$ dB, $Q_{\mathrm{ext}} = 100$, $B = 100$ Hz, will be

$$(\Delta f_{\mathrm{rms}})^{\mathrm{prim}} = \frac{6 \times 10^9}{10^2} \cdot \left(\frac{1.38 \times 10^{-23} \times 2.9 \times 10^2 \times 10^2 \times 3.16 \times 10^2}{10^{-2}}\right)^{1/2} = 6.74\ \mathrm{Hz}.$$

The total noise in a 100-Hz band and for $Q_{\mathrm{ext}} = 100$ is then

$$\Delta f_{\mathrm{rms}} = [(\Delta^2 f_{\mathrm{rms}})^{\mathrm{prim}} + (\Delta^2 f_{\mathrm{rms}})^{\mathrm{mod}}]^{1/2} = 18\ \mathrm{Hz}.$$

This constitutes a significant increase over the expected primary noise. If the voltage supply is used, the effect of its own noise (Fig. 4) should be negligible as it is much lower than the diode noise at 30 kHz and, therefore, the total FM noise should be close to the above figure.

In order to eliminate the influence of up-conversion on FM noise, a current bias source must be used. However, as can be seen from Fig. 4, the noise of our current supply comes quite close to that of the diode at 30 kHz, if no filter is used. One can therefore expect almost the same amount of up-converted noise as in the previous case. If the filter indicated in Fig. 4 is used in conjunction with this current source, this up-conversion will be eliminated and the resultant FM noise should be close to the primary noise. This measure will be needed even more at $I_0 = 10$ mA where the influence of up-conversion should be more pronounced since the ac diode resistance is smaller. Conversely, at 20 mA, there should be less up-converted noise at $I_0 = 20$ mA because of higher ac diode resistance. In this latter case the bias supply noise will contribute more strongly if the current source is used without a filter.

These preliminary conclusions will be confronted with actually measured values of AM and FM noise in the next section. One additional remark is in order at this point; the values of M_{AM} and M_{FM} seem to have been chosen somewhat arbitrarily for the primary noise calculations. This is really so to an extent since otherwise they would have to be determined from the relatively involved theories in [5] or [6], which would also require a suitable large-signal analysis of the BARITT diode. However, it has been observed with IMPATT's [6] that the AM noise is always smaller than would correspond to the simple noise measure M, i.e., $M_{\mathrm{AM}} < M$ and, consequently $M_{\mathrm{FM}} > M$. Assuming that this is also true for BARITT's (although on a smaller scale because of less pronounced nonlinearities), and taking $M = 23$, we estimated $M_{\mathrm{AM}} = 20$ and $M_{\mathrm{FM}} = 25$. The experimental results in the next section will show whether this estimate was reasonable.

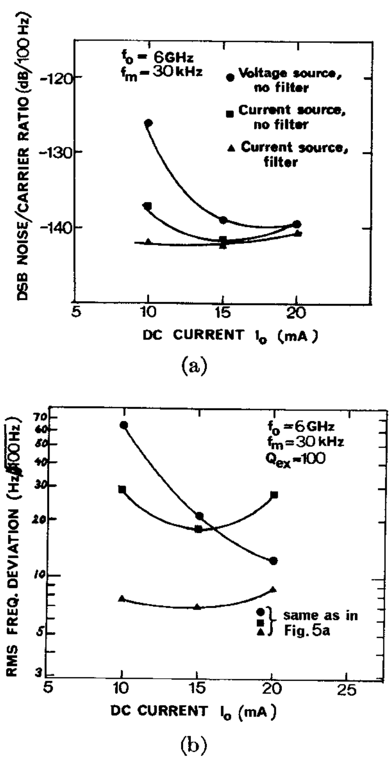

Fig. 5. (a) AM noise as a function of bias current. (RF power adjusted to maximum for each value of bias current.) (b) FM noise as a function of bias current. (RF power adjusted to maximum for each value of bias current.)

IV. AM AND FM NOISE: MEASUREMENT RESULTS AND DISCUSSION

The AM and FM noise of our BARITT diode oscillator was measured by means of a direct-detection noise-measuring system described in detail in [15]. It is basically the system of Ashley *et al.* [16] with some modifications which improved sensitivity at low modulation frequencies.

The BARITT diode, the data of which are given in Table I, was imbedded in a J-band coaxial resonator with two tuning slugs which made possible fine adjustment of frequency and of output coupling. The diode was operated at 6 GHz over the bias current range 10–20 mA. The maximum power achieved was 10 mW. Both voltage and current bias sources were used and their noise properties measured with the results indicated in Fig. 4. The noise measurements were carried out in the baseband frequency range 100 Hz–50 kHz.

In Fig. 5, the AM and FM noise at 30 kHz off the carrier is shown in dependence on bias current for the three different bias sources indicated. At 15 mA, the increase in AM noise due to up-conversion (when changing from current source to voltage source) of low-frequency diode noise is approximately 3 dB, while at 10 mA the change is approximately 15 dB, and at 20 mA it is only 1 dB, which is obviously within the limits of measurement inaccuracy. These values are in good agreement with the calculated ones in Section III. This is also true for the measured and estimated values of primary noise.

The FM noise, when using the filtered current source,

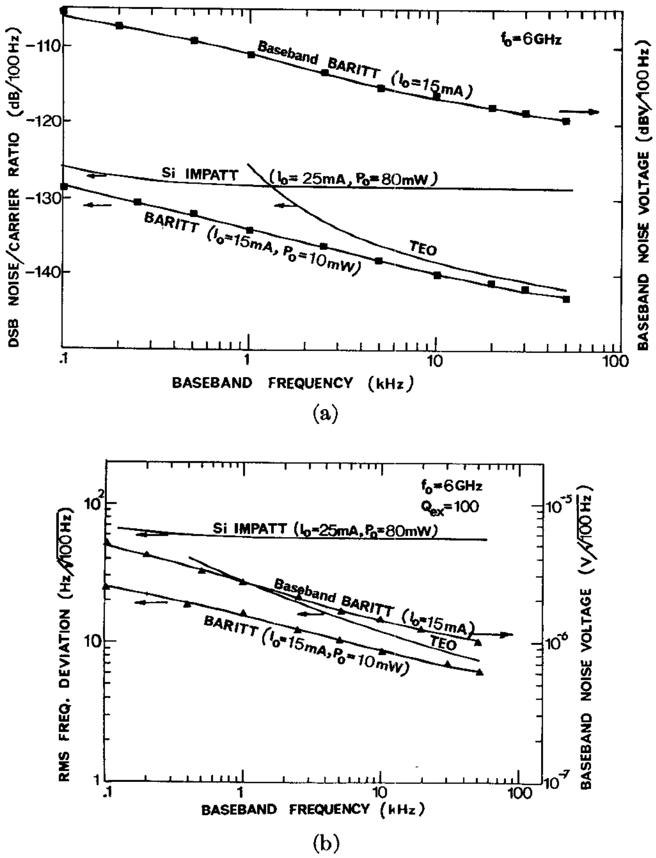

Fig. 6. (a) AM noise of Si IMPATT, Gunn, and BARITT oscillators in dependence on baseband frequency (distance from carrier). Also shown is the low-frequency BARITT diode noise voltage spectrum. (b) FM noise of Si IMPATT, Gunn, and BARITT oscillators in dependence on baseband frequency (distance from carrier). Also shown is the low-frequency BARITT diode noise voltage spectrum.

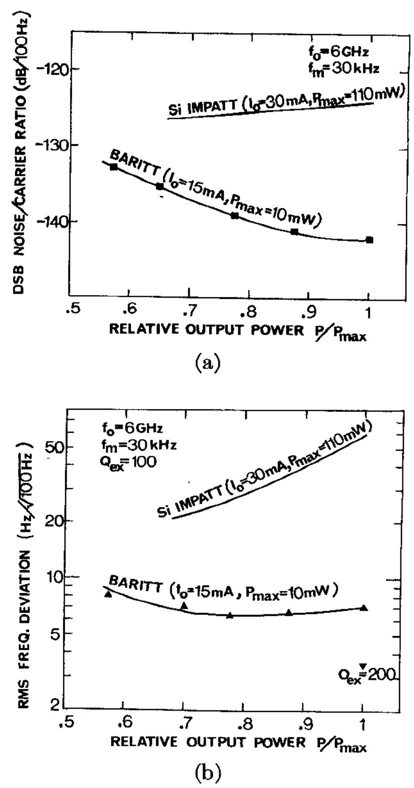

Fig. 7. (a) AM noise of Si IMPATT and BARITT oscillators as a function of the relative output power P/P_{max}. (b) FM noise of Si IMPATT and BARITT oscillators as a function of the relative output power P/P_{max}.

was measured to be 7 Hz/(100 Hz)$^{1/2}$ at 15-mA dc current and with Q_{ext} = 100, which agrees with the estimated value of FM primary noise. This seems to confirm that the value of M_{FM} = 25 dB was well chosen. The actual M_{FM} for pure white noise may be somewhat lower however, since at 30 kHz we are still in the reign of flicker noise. The same reasoning applies to the above mentioned AM noise. More importantly, there is a good agreement between the estimated and measured values of the up-converted FM noise at 15-mA dc current; Δf_{rms} jumps from 7 Hz/(100 Hz)$^{1/2}$ to 21/(100 Hz)$^{1/2}$ when changing from filtered current source to voltage source, which compares well with the calculated value of 18 Hz/(100 Hz)$^{1/2}$. Fig. 5(a) and (b) also shows that while additional filtering of the current supply does not influence AM noise too much, its effect on FM noise is quite pronounced and very desirable. Only by using this filter could the FM noise be reduced to the primary level.

It is known that BARITT diodes exhibit a certain amount of "flicker" noise at frequencies below approximately 100 kHz [11]. Fig. 6(a) and (b) shows the results of the AM and FM noise measurements in the baseband frequency range 100 Hz–50 kHz, taken with the diode described by Table I under "ideal" conditions, i.e., with all modulation noise (both due to the diode noise and bias supply noise) eliminated. Also, the diode was operated at I_0 = 15 mA where minimum "primary noise" was obtained. A comparison is made with the baseband noise voltage. It is seen that in both cases the two curves run almost parallel, indicating a definite correlation between the baseband noise voltage and the AM and FM noise, even if all modulation is suppressed. This can be only explained by referring to (8). The power spectral density $S_{FF}(\Omega)$ at baseband frequencies, of the intrinsic noise process, can significantly influence the resultant noise measure at the frequency of oscillation. Hence any intrinsic flicker noise, superimposed on the so far known white noise sources in BARITT's (shot noise and velocity fluctuation noise) can be up-converted in a manner which is partially immune against the external circuitry. Therefore, this kind of "internal" up-conversion can be eliminated only by removing the flicker noise source itself; this will probably be the matter of improving device technology.

Also shown in Fig. 6(a) and (b) are the data measured by the author on a 6-GHz Si IMPATT, and those of a typical X-band TEO, taken from [17] and scaled to 6 HGz and Q_{ext} = 100. It is seen that both the AM and FM noise of the BARITT compare most favorably with those of its most serious contender, the TEO (Gunn).

Finally, Fig. 7(a) and (b) shows the dependence of the BARITT AM and FM noise on RF power level. This measurement was again taken under conditions giving lowest noise, i.e., with the filtered current supply and at $I_0 = 15$ mA. Previous investigation of IMPATT diodes [6],[7] showed both theoretically and experimentally that their FM noise could be significantly reduced by reducing (via undercoupling) the output power level by some 30 percent. Fig. 7 shows that, in the case of BARITT's, nothing can be gained with this approach as the change of $\Delta f_{\rm rms}$ with power level is very small in the high power region, compared to IMPATT's. Moreover, the AM noise would deteriorate, instead of slightly improve as is the case with IMPATT's. Both these effects are related to the fact that the noise measure of BARITT's, unlike that of IMPATT's, does not increase very much with RF power level. No detailed information is available for TEO's in this respect, but from [17] one can deduce a behavior similar to that of BARITT's.

It is seen, however, that even at full RF power level the BARITT oscillator noise can be very low indeed: at 30 kHz off the carrier the AM noise is some 142 dB/100 Hz below the carrier and the rms frequency deviation is approximately 7 Hz/(100 Hz)$^{1/2}$ for $Q_{\rm ext} = 100$, or 3.5 Hz/(100 Hz)$^{1/2}$ for $Q_{\rm ext} = 200$. The latter figure is very close to the typical requirement on local oscillators in multihop FM systems [18]. Therefore, no special high-Q cavity would be needed from the short-term stability point of view. Nevertheless, such a cavity might have to be employed because of the BARITT's sensitivity to temperature [19] and resulting long-term stability problems. Alternatively, the modulation capability of the BARITT diode could be utilized for long-term stabilization in an automatic-frequency-control loop in a similar way as it was done with an unspecified voltage-controlled oscillator in [20].

V. SUMMARY AND CONCLUSIONS

AM, FM, and baseband noise measurements on a BARITT diode oscillator were carried out under various operating conditions. These data were found to be in good agreement with the calculated primary and modulation (up-converted) noise. The results can be summarized as follows.

1) Depending on the value of bias current, both AM and FM up-conversion noise can be significant when compared to the primary noise. At higher bias currents the effect of up-conversion on AM noise is smaller than that on FM noise. This is in contrast to IMPATT oscillators where FM up-conversion noise was found to be negligible.

2) Because of the generally low diode noise level, the AM and FM noise of BARITT's can be significantly degraded even by the up-conversion of the baseband bias supply noise. To remove up-conversion by modulation completely, a well-filtered high-impedance bias supply should be used.

3) Even if all the modulation is suppressed, the AM and FM noise of BARITT's has a "flicker" noise component at frequencies below approximately 100 kHz, running almost parallel with the baseband flicker noise curve. This can only be caused by internal conversion in the diode due to the combined effect of the diode nonlinearity and large-signal excitation; the external circuit has minimum influence on it.

4) The AM noise continually decreases with RF power level whereas the FM noise exhibits a minimum and a slight increase afterwards. However, this latter effect is so small that no significant noise reduction can be obtained by reducing the RF power level, which again is in contrast to the behavior of IMPATT's [6],[7].

5) In the range of baseband frequencies 100 Hz–50 kHz, the BARITT compares very favorably with the TEO and IMPATT. Even at 100 Hz off the carrier, the 6-GHz BARITT oscillator had lower FM noise than a typical 6-GHz IMPATT, while the TEO is known to sometimes have FM noise higher than IMPATT's at such low frequencies.

The typical performance at 30 kHz off the carrier, of the BARITT oscillator tested, was AM noise of some -142 dB/100 Hz and FM noise of approximately 3.5 Hz/(100 Hz)$^{1/2}$ with $Q_{\rm ext} = 200$. These values clearly show, at least from the short-term stability point of view, the potential suitability of BARITT's for local oscillator applications. Also, in our limited observation, the p^+-n-p^+ BARITT seems to be a much "tougher" device compared to the TEO as far as their sensitivity to various circuit transients is concerned. One problem in the above application may be that of temperature sensitivity and long-term stability in general. This area requires further study and experimentation; the outcome will probably depend on the feasibility of employing these otherwise attractive devices as local oscillators in microwave radio communication systems.

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REFERENCES

[1] H. A. Haus, H. Statz, and R. A. Pucel, "Noise measure of metal-semiconductor-metal Schottky-barrier microwave diodes," *Electron. Lett.*, vol. 7, pp. 667–668, Nov. 1971.

[2] ——, "Velocity fluctuation noise in metal–semiconductor–metal diodes," *Proc. IEEE* (Lett.), vol. 60, pp. 644–645, May 1972.

[3] A. Sjölund, "Small-signal noise analysis of p^+-n-p^+ BARITT diodes," *Electron Lett.*, vol. 9, pp. 2–4, Jan. 1973.

[4] R. L. Kuvås, "Noise in IMPATT diodes: Intrinsic properties," *IEEE Trans. Electron Devices*, vol. ED-19, pp. 220–233, Feb. 1972.

[5] ——, "Noise in single-frequency oscillators and amplifiers," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-21, pp. 127–134, Mar. 1973.

[6] J. L. Fikart, "A theory of oscillator noise and its application to IMPATT diode oscillators," Ph.D. dissertation, Univ. Alberta, Edmonton, Alta., Canada, 1973.

[7] J. L. Fikart and P. A. Goud, "A theory of oscillator noise and its application to IMPATT diodes," *J. Appl. Phys.*, vol. 44, pp. 2284–2296, May 1973.

[8] D. J. Coleman and S. M. Sze, "A low noise metal-semiconductor-metal (MSM) microwave oscillator," *Bell Syst. Tech. J.*, vol. 50, pp. 1695–1699, May–June 1971.
[9] C. A. Lee and G. C. Dalman, "Local oscillator noise in a silicon Pt-n-p$^+$ microwave diode source," *Electron Lett.*, vol. 7, pp. 565–566, Sept. 1971.
[10] G. Björkman and C. P. Snapp, "Small-signal noise behavior of companion p$^+$-n-p$^+$ and p$^+$-n-ν-p$^+$ punchthrough microwave diodes," *Electron Lett.*, vol. 8, pp. 501–503, Oct. 1972.
[11] J. Helmcke, H. Herbst, M. Claassen, and W. Harth, "FM noise and bias-current fluctuations of a silicon Pd-n-p$^+$ microwave oscillator," *Electron Lett.*, vol. 8, pp. 158–159, Mar. 1972.
[12] H. Herbst and W. Harth, "Frequency modulation sensitivity and frequency pushing factor of a Pd-n-p$^+$ punchthrough microwave diode," *Electron Lett.*, vol. 8, pp. 358–359, July 1972.
[13] W. A. Edson, "Noise in oscillators," *Proc. IRE*, vol. 48, pp. 1454–1466, Aug. 1960.
[14] K. Kurokawa, "Some basic characteristics of broadband negative resistance oscillator circuits," *Bell Syst. Tech. J.*, vol. 48, pp. 1937–1955, July 1969.
[15] J. L. Fikart and P. A. Goud, "The direct-detection noise-measuring system and its threshold," *IEEE Trans. Instrum. Meas.*, vol. IM-21, pp. 219–224, Aug. 1972.
[16] J. R. Ashley, C. B. Searles, and F. M. Palka, "The measurement of oscillator noise at microwave frequencies," *IEEE Trans. Microwave Theory Tech.* (*Special Issue on Noise*), vol. MTT-16, pp. 753–760, Sept. 1968.
[17] P. A. Levine, H. C. Huang, and H. Johnson, "IMPATTs shoot for Gunn noise levels," *Microwaves*, pp. 52–56, Apr. 1972.
[18] D. B. Leeson, "Short term stable microwave sources," *Microwave J.*, vol. 13, pp. 59–69, June 1970.
[19] M. El-Gabaly, J. Nigrin, and P. A. Goud, "Stationary charge transport in metal-semiconductor-metal (MSM) structures," *J. Appl. Phys.*, to be published.
[20] Raytheon Company, "60 GHz line of sight microwave link," *Microwave J.*, vol. 12, p. 38, Nov. 1969.

New Theory and Design for Hairpin-Line Filters

ULRICH H. GYSEL, MEMBER, IEEE

***Abstract*—Hairpin-line and hybrid hairpin-line/half-wave parallel-coupled-line filters are preferred filters for microstrip and TEM printed-circuit realizations. This class of filters offers small size and, in general, needs no ground connections for resonators.**

A new design theory is presented that is based on a sparse capacitance matrix for the array of coupled lines that constitute the filter, as opposed to a sparse-inductance-matrix assumption in previous theories that is much harder to satisfy. It is shown that to a good approximation, hairpin-line filters result from frequency-scaling half-wave parallel-coupled-line filters. Because of this, the bandwidth can be accurately predicted.

Design procedures are given for Type-*A* filters, which are useful up to 20-percent bandwidth. A variety of hybrid hairpin-line/half-wave parallel-coupled-line filters is possible, and their design is explained. Numerical results for a number of designs and experimental results for a 5-percent bandwidth filter are included.

I. INTRODUCTION

THE hairpin-line filter, like the half-wave parallel-coupled-line filter, is one of the preferred configurations in stripline or microstrip because ground connections are not required. Basically, the hairpin-line filter can be thought of as a folded version of a half-wave parallel-coupled-line filter. The hairpin-line filter makes a much more compact filter than the half-wave parallel-coupled-line filter, but is expected to have much the same performance. However, the additional coupling between the two lines that constitute a hairpin resonator complicates the design.

The image parameters for the infinite periodic hairpin line have been reported by Bolljahn and Matthaei [1]. Only recently, design equations have been given for finite length hairpin-line filters by Cristal and Frankel [2]. Their design theory is based on the assumption of a sparse *inductance* matrix for the array of coupled lines. But for most parallel-coupled-line-type filters, the assumption of a sparse *capacitance* matrix is made. This, as Cristal and Frankel [2] state, corresponds much more closely to the physical reality than does neglecting inductive coupling beyond nearest neighbors. Further, this design theory [2] needs an empirically determined bandwidth contraction factor, depending on the hairpin resonator coupling.

The present paper gives exact equivalent circuits for odd-order hairpin-line filters of Types *A* and *B* that are based on a sparse capacitance matrix. This leads to designs that are exact for any practical purposes up to 20-percent bandwidth for Type-*A* filters and up to 50-percent bandwidth for Type-*B* filters. In particular, the bandwidth can be predicted accurately. From a theoretical point of view, it is most interesting that the new design method explains the bandwidth contraction factor in [2] and particularly shows that it is independent of the number of resonators, of passband ripple, and, largely, of bandwidth. This is the

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The author is with the Stanford Research Institute, Menlo Park, Calif. 94025.