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Design of Stable, Very Low Noise, Cavity-Stabilized IMPATT Oscillators for *C* Band

BERNARD F. VAN DER HEYDEN

Abstract—**Two types of *C*-band IMPATT oscillators, which easily meet the noise and stability requirements for use as local oscillators in microwave FM communications equipment, are described. Both types use a transmission cavity-stabilization circuit as proposed by Kurokawa. In one of them a TE_{103} mode rectangular invar cavity is used for stabilization, while in the other the coupling is made via a high-*Q* cylindrical TE_{011} mode cavity.**

Although the Si IMPATT diode is inherently noisy, it is shown that a proper choice of circuit parameters and diode characteristics leads to measured FM noise levels of less than 0.2 Hz in a 100-Hz band.

With respect to frequency stability, special attention is paid to hysteresis-free compensation of temperature effects and to the influence of changes with time and ambient temperature of the diode and of the internal atmosphere of the cavity. By careful processing and sealing, an average temperature stability of better than -0.4 ppm/°C was realized with temperature cycling between 26 and 51°C over a period of 450 h.

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The author is with the Product Division Elcoma, Philips Development Laboratories for Transmitting Tubes and Microwave Components, Eindhoven, The Netherlands.

I. Introduction

MICROWAVE signal sources with a high degree of short- and long-term frequency stability find application in systems for radar and FM communication (telephone, television). In many of these systems the local oscillator consists of a crystal-controlled transistor oscillator, followed by a wide-band frequency multiplier, in order to obtain an output signal at the desired microwave frequency. A requirement for the FM noise of the oscillator is set by the CCIR recommendations, which, for example, for an 1800-channel telephone microwave link specify that, measured in a 3.1-kHz band at baseband frequencies from 10 kHz to 10 MHz, the noise level is 80 dB below the level of a 140-kHz rms deviation test tone. If we suppose no preemphasis and no psophometric weighting, this implies a single-sideband FM noise power $\leq$ 10 pW referred to a 1 mW test tone level, or expressed in other units, an rms frequency deviation $\Delta f_{rms} \leq 2.5$ Hz, measured in 100 Hz in the baseband frequency range indicated. Additional requirements are tunability, freedom of microphonic effects and, in general, an output signal level of at most 50 mW, which is free from harmonics and spurious frequencies. Long-term variations of frequency, in particular those caused by temperature change, are required to be within $\pm$ 10 ppm for ambient temperatures from 0 to 50°C.

At present, however, interest is growing for solid-state devices which directly generate the signal at the desired frequency, and which still meet the strict requirements with respect to FM noise and long-term frequency stability. These devices are less complex and can be manufactured possibly at lower cost. Various oscillators, equipped with either Gunn or IMPATT diodes, have been reported, employing high-Q cavity stabilization for long-term stability and FM noise reduction. In a considerable number of these, the Gunn diode is used because of its lower inherent noise level compared to that of the IMPATT diode [1]. However, in television and satellite communications equipment, where low noise near the carrier is required, and in applications where a higher output power is desirable, the Gunn diode is less attractive ($1/f$ noise, low power) and the IMPATT diode becomes in favor. Moreover, Gunn diode parameters are more temperature dependent than those of the IMPATT diode.

Many existing Gunn oscillators employ reaction cavity stabilization. It is known, however, that these oscillators may be subject to instabilities such as mode jumping and generation of unwanted frequencies.

Ashley and Searles [8] already described the microwave oscillator noise reduction by a transmission stabilizing cavity, with data taken on klystrons and an IMPATT diode oscillator.

The purpose of the present paper is to discuss the design and technology of two types of oscillators equipped with IMPATT diodes for operation at frequencies around 7 GHz, and to present experimental results which indicate that severe stability requirements can be met, and that the noise level is even lower than that of usual Gunn oscillators. The oscillators are based on a circuit using transmission cavity stabilization as described by Kurokawa and Magelhaes [2] and patented by Harkless [7], whose circuit is known for its stable operation because of the off-resonance decoupling between the diode circuit and the load. In both types (Figs. 2 and 3) the diode is mounted at the end of a coaxial line, which is terminated in its characteristic impedance Z_0 and coupled to the high-Q transmission cavity. In the first design a TE_{103} mode rectangular cavity is used (R circuit), in the second a TE_{011} mode cylindrical cavity with a much higher unloaded Q value.

Special attention will be paid to the main factors that influence the long-term frequency stability, as there are temperature- and time-dependent changes of the circuit (thermal expansion, hysteresis, atmosphere within the cavity) and of the diode. Practical solutions are given to ensure this stability, this being perhaps the most critical point in the design of solid-state oscillators for the purpose envisaged.

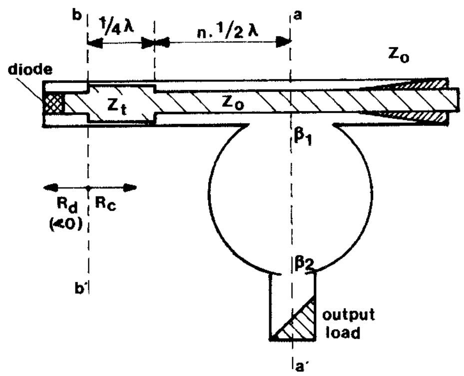

Fig. 1. Schematic drawing of a transmission cavity-stabilized oscillator.

II. Electrical Design Considerations

A schematic drawing of the Kurokawa circuit is given in Fig. 1. We define a reference plane aa' in the coaxial line through about the center of the coupling to the cavity, chosen such that the load impedance at cavity resonance, as seen to the right, has a real value. This impedance is transformed down by a $\frac{1}{4}\lambda$ transformer with characteristic impedance Z_t to a value R_c in reference plane bb'. Of course, an additional line length of $n \cdot \frac{1}{2}\lambda$ may be present. Tjassens [4] introduced the input coupling factor from the coaxial line to the cavity β_1 and the output coupling factor β_2, in order to derive expressions for R_c and the circuit efficiency η_c, defined as the ratio of the useful power dissipated in the load to the power delivered to the circuit at aa'. If Q_0 represents the unloaded Q of the cavity and

- Q_{e1} external Q, due to losses in the input circuit with both sides of the coaxial line matched;
- Q_{e2} external Q, due to losses in the output circuit; then the coupling factors are defined by

$$\beta_1 = Q_0/Q_{e1} \quad \beta_2 = Q_0/Q_{e2}.$$

The basic design now involves the determination of β_1 and β_2 in accordance with the following requirements.

1) To fulfil the oscillation condition, R_c has to be equal to the magnitude of the negative resistance R_d, as seen in reference plane bb' in the direction of the generator (diode) part of the circuit, which is also supposed to be tuned to the cavity resonance frequency. Then,

$$R_c = \frac{1 + \beta_2}{1 + 2\beta_1 + \beta_2} \cdot \frac{Z_t^2}{Z_0} = -R_d. \tag{1}$$

The value of R_d must correspond to that operating point of the diode, at which the inherent FM noise level is minimum for a given dc diode current. As is known, this operating point, found from noise measurements in a low-Q Iglesias-type coaxial oscillator [3], is at an impedance level higher than that corresponding to maximum available power from the diode.

2) The circuit efficiency must be acceptable, and is shown in [4] to be

$$\eta_c = \frac{2\beta_1\beta_2}{(1 + \beta_2)(1 + 2\beta_1 + \beta_2)}. \tag{2}$$

3) The circuit stabilization factor S must be as large as possible. This factor is defined as follows as the ratio of the frequency derivatives of the susceptances B at resonance, when looking in reference plane bb' into the cavity and the generator parts of the circuit, respectively.

$$S = \left(\frac{\partial B}{\partial \omega}\right)_c \bigg/ \left(\frac{\partial B}{\partial \omega}\right)_g. \tag{3}$$

It can be derived that

$$\left(\frac{\partial B}{\partial \omega}\right)_c = \frac{-4\beta_1 Q_0}{\omega_0(1+\beta_2)^2 Z_0} \cdot \left(\frac{Z_0}{Z_t}\right)^2. \tag{4}$$

Combination of (1), (2), and (4) gives

$$\left(\frac{\partial B}{\partial \omega}\right)_c = \frac{2Q_0}{\omega_0 R_d} \cdot \frac{\eta_c}{\beta_2}.$$

For the generator part of the circuit, it is easily found that

$$\left(\frac{\partial B}{\partial \omega}\right)_g = \frac{2Q_{0g}}{\omega_0 R_d}$$

where Q_0 and Q_{0g} (< 0) are the unloaded Q values for the cavity and generator (diode) parts of the circuit. Hence

$$S = \left|\frac{Q_0}{Q_{0g}}\right| \cdot \frac{\eta_c}{\beta_2} = \left|\frac{Q_0}{Q_{0g}}\right| \cdot \frac{2\beta_1}{(1+\beta_2)(1+2\beta_1+\beta_2)}. \tag{5}$$

It is concluded that Q_0 should be high, of course, combined with a tight input coupling and a weak output coupling, still giving a reasonable circuit efficiency. From (2), however, it follows that for a certain value of η_c required, we must fulfill the condition

$$\beta_2 > \frac{\eta_c}{1-\eta_c}$$

so that always

$$S < \left|\frac{Q_0}{Q_{0g}}\right| \cdot (1-\eta_c). \tag{6}$$

Thus high efficiency and a high degree of stabilization are contradictory requirements.

An acceptable value for η_c is 40 percent, which means $\beta_2 > \frac{2}{3}$. Taking $\beta_2 = 0.75$ gives $\beta_1 \approx 12$, while the required value of Z_t is then determined by (1). In the actual designs, however, Z_t was taken equal to Z_0 and β_1 modified in such a way as to fulfill condition (1), because low values for Z_t involve narrow spacings and, consequently, these impose constructional difficulties to ensure stability under conditions of shock, vibration, or varying ambient temperature.

III. Description of Oscillator Circuits

A. *R Circuit* (Fig. 2)

In order to minimize the frequency shift at varying ambient temperature, the cavity is made from specially heat-treated silver-plated invar, with a stable thermal expansion of about $1.5 \times 10^{-6}/°C$. The cavity is $1\frac{1}{2}\ \lambda_g$ long at the highest tuning frequency, corresponding to a theoretical

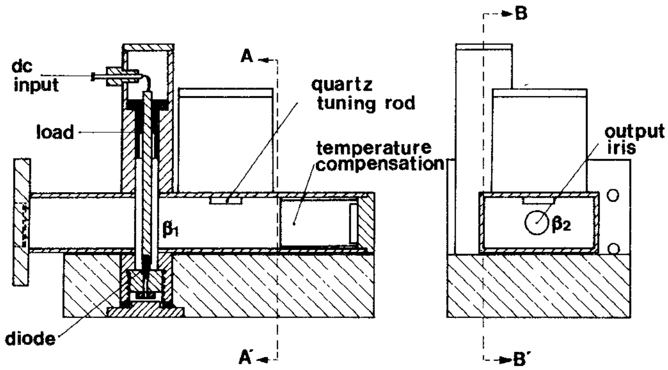

Fig. 2. IMPATT oscillator, stabilized with a TE_{103} mode waveguide cavity (R circuit).

$Q_0 \approx 13\,000$, and is provided with a circular output iris covered by a glass window. It matches standard WG 14 waveguide with the correct value of β_2.

The packaged diode used is of the Si p^+-n-n^+ type, having a junction capacitance of 0.25 pF at a breakdown voltage of about 108 V. A 50-Ω coaxial line, made from copper, is coupled to the cavity on the small sidewall, halfway along the cavity where the z component of the magnetic field is maximum. One side of the line carries the diode, clamped in a copper-to-aluminium heat sink block, at a distance of approximately $\frac{1}{4}\lambda$ to midheight of the cavity. This minimum distance was chosen because then the largest bandwidth is obtained when tuning the cavity. The other side of the line is terminated in a short-stepped load, machined from Eccosorb, showing a VSWR < 1.05 within a 4–8-GHz frequency band. As this material is dc isolating, bias current to the diode can be easily supplied.

Capacitive tuning is accomplished by a quartz rod, adjustable within an invar housing, which gives a cavity tuning range of 400 MHz. This design ensures smooth low-loss tuning, thus the highest possible Q. The power output can be optimized at a desired frequency by adjusting the distance between diode and cavity with the aid of thin insertion rings.

The thermal expansion of invar (α_i) leading to an equal relative decrease of the resonant frequency with increasing temperature, is corrected for by a built-in compensation at a position shown in Fig. 2. The compensator is made from copper, hence a material with an appreciable larger thermal expansion coefficient α_m. For ideal compensation at midband wavelength λ_m, its length p can be calculated from

$$p = \frac{3\lambda_m}{2\left(\frac{\alpha_m}{\alpha_i}-1\right)\left[1-\left(\frac{\lambda_m}{2a}\right)^2\right]^{3/2}} \tag{7}$$

where a is the waveguide width.

Structures of this kind, often described in a way by which a block of material is brazed with one side to the invar endplate, have the great disadvantage, however, that with varying temperature hysteresis phenomena are introduced due to plastic deformation and bending effects at the large brazing interface (bimetal action). In the present construction, this effect has been eliminated by brazing the

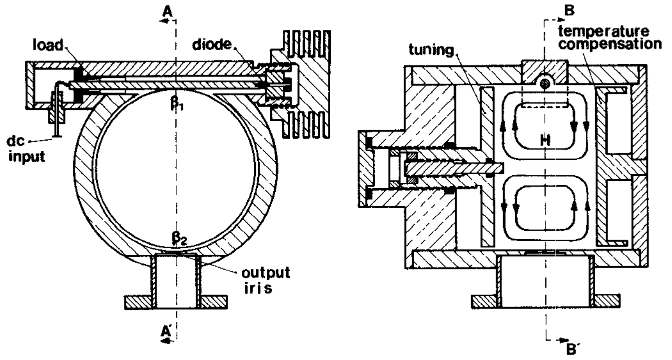

Fig. 3. IMPATT oscillator, stabilized with a TE_{011} mode cylindrical cavity (C circuit).

compensator with small, somewhat flexible, ridges to the cavity walls.

When discussing frequency stabilities in the order of 1 ppm/°C, the change of diode impedance with temperature becomes equally important as thermal expansion of the cavity material, and with it, some overcompensation is necessary. From this point of view it seems useless to look for cavity-materials with still smaller thermal expansion. As a change of the dielectric medium inside the cavity has a large effect on df/dT, the system has to be filled with dry air or preferably nitrogen, after which it is completely sealed. This point is discussed more in detail in Section IV.

B. *C Circuit* (Fig. 3)

In this design microwave power is coupled from the coaxial circuit into a cylindrical cavity, made from silver-plated invar, in such a way that the most favorable conditions exist for the excitation of the TE_{011} mode, this being the mode with the highest Q at the smallest volume. Proper selection of the operating area in the mode chart, to minimize the number of interfering and crossing modes, gives a cavity diameter-to-length ratio of 1.6, which, at a midband frequency of 6.75 GHz, results in a diameter of 6.5 cm. The theoretical unloaded Q value is then 34 000.

A number of other modes are possible, most of them lying outside the frequency range, where the oscillation condition is fulfilled. Because all TM modes need an RF current component flowing from the end plates to the cylindrical surface, these are suppressed by designing the flat endplates as noncontacting piston-shaped disks. One of them serves as a tuning element, provided with a fine-tuning pin in the center, while the other is the temperature compensator, made from copper in a shape to minimize hysteresis effects. For constructional reasons, the distance between the diode and the input coupling hole has to be in the order of $\frac{3}{4}\lambda$, which restricts the tuning bandwidth to approximately 300 MHz for a fixed diode position. This may, however, be an advantage, because the nearest interfering mode, the TE_{311}, occurs at a frequency about 450 MHz higher, and so needs no extra suppression. The degenerate TM_{111} mode associated with the TE_{011} mode, often leading to lower Q values, was not detectable due to the presence of the gaps and accurate machining. Making one of the disks concave, as reported elsewhere [8], was not necessary. For the remaining parts of the circuit, the same remarks hold as given in Section III-A.

Special attention must be paid to proper heat treatment of the invar, used for both cavities. A stable and low thermal expansion is obtained by first heating the rough material at 830°C, followed by a quenching procedure for minimum α_i; after machining, annealing at 315°C removes the stresses and, finally, aging at 95°C for 48 h stabilizes the material [6].

IV. Frequency Stability

The short-term frequency stability, as characterized by the output FM noise, is determined by the inherent noise of the diode, reduced by the circuit stabilization factor S (see also Section V). Long-term and temperature-dependent frequency stability is governed by changes of the diode impedance and resonant frequency of the stabilizing circuit with time and ambient temperature. The latter subject will be discussed in more detail.

The changes in time of the cavity dimensions can be made sufficiently small by the heat treatment of the invar, while the influence of thermal expansion can be compensated in a way described in Section III.

Changes of the medium inside the cavity and the variation of its dielectric constant with temperature play a very important role. Montgomery [5] has given an expression for the dielectric constant ε of humid air as a function of the partial pressures P_a and P_w (in millimeters of mercury) of the air and water vapor, respectively, and the absolute temperature T. The relative change of the cavity resonant frequency is then $\Delta f/f \approx -\frac{1}{2}\cdot\Delta\varepsilon$, and now we shall consider two cases.

A. *Open Cavity with $P = P_a + P_w =$ Constant*

After introduction of the relative humidity η and a formula for the saturated water vapor pressure derived from Clapeyron's law,

$$P_{w_{\text{sat}}} = 1.26\cdot 10^9 e^{-5300/T} = 1.26\cdot 10^9 T\phi(T) \text{ mmHg}$$

where

$$\phi(T) = \frac{e^{-5300/T}}{T} \tag{8}$$

we can derive for the frequency change with temperature

$$\frac{\Delta f}{f} = \frac{1}{T^2}\left[\left(1.05\cdot 10^{-4}P + \frac{P_w}{T}\right)\cdot \Delta T - \frac{1}{2}\Delta P_w\right] \tag{9}$$

or

$$\frac{1}{f}\frac{\Delta f}{\Delta T} = \frac{1}{T^2}\left[1.05\cdot 10^{-4}P + 1.26\cdot 10^9 \eta\phi(T) - 3.34\cdot 10^{12}\eta\frac{\phi(T)}{T} - 6.3\cdot 10^8 \frac{\Delta\eta}{\Delta T}\cdot T\cdot\phi(T)\right]. \tag{10}$$

To give some numerical examples, if T remains constant, (9) reduces to

$$\frac{\Delta f}{f} = -6.3\cdot 10^8 \frac{\phi(T)}{T}\cdot\Delta\eta.$$

This means for $\Delta\eta = 1$ percent at $T = 298$ K: $\Delta f/f = -1.3$ ppm. If, however, T increases at constant P_w, it is found from (9) that for $T = 298$ K, $P = 760$ mmHg and an initial humidity $\eta_i \doteq 0.60$: $(1/f) \cdot (\Delta f/\Delta T) = +1.4$ ppm/°C. At constant temperature, an air pressure variation $\Delta P = +10$ mmHg would give a frequency shift $\Delta f/f = -3.5$ ppm. These are all rather large effects.

B. Sealed Cavity

In the case of a cavity, initially filled with air of humidity η_i at a temperature T_i, and subsequently sealed, we apply the ordinary gas laws and so long as the temperature remains above the dewpoint, we can derive that the frequency shift is given by

$$\left(\frac{\Delta f}{f}\right)_{T>Td} = 6.3 \cdot 10^8 \eta_i \left(\frac{1}{T_i} - \frac{1}{T}\right) \cdot \phi(T_i)$$

or

$$\left(\frac{1}{f}\frac{\Delta f}{\Delta T}\right)_{T>T_d} \approx 6.3 \cdot 10^8 \eta_i \frac{\phi(T_i)}{T_i^2}. \tag{11}$$

At $T_i = 298$ K and $\eta_i = 0.60$ this means a variation of 0.27 ppm/°C, which is about five times smaller than the value found for the unsealed cavity. If the temperature drops below the dewpoint T_d, where $\eta = 1$, water precipitates, and very large frequency shifts result. The value of T_d can then be determined by graphical solution from

$$\phi(T_d) = \eta_i \phi(T_i).$$

It can be shown that below the dewpoint, Δf is given by

$$\left(\frac{\Delta f}{f}\right)_{T<T_d} = 1.13 \cdot 10^5 \cdot \left[\left(1 + \frac{5580}{T_d}\right) \cdot \phi(T_d) - \left(1 + \frac{5580}{T}\right) \cdot \phi(T)\right] \tag{12}$$

which, under the same initial conditions, means -4.4 ppm/°C, that is, 16 times as large!

According to the formulas given, Fig. 4 represents the resonant frequency shift as a function of temperature for the unsealed and sealed cavity at different initial humidities η_i. The fact clearly emerges that for operation between 0 and 50°C, the initial humidity must be certainly less than 20 percent, and that for small df/dT the cavity must be filled with dry air and has to be completely sealed. An inert gas such as nitrogen is preferred to prevent changes in composition and to maintain a good cavity surface finish for a long time. There is still another reason for requiring a completely dry atmosphere. If any moisture is present, then, dependent on temperature fall or rise, part of the water can be absorbed or released again slowly by metal surfaces or built-in organic materials such as plastics. This causes the frequency to creep exponentially to its final value. A calculation shows that even an amount as small as 0.1 mg of water entering into the atmosphere within the R cavity results in a frequency shift of -13 ppm! The necessity to make the whole system vacuum tight is clear. Where possible, metal parts are brazed, and the complete oscillator is outgassed in vacuum. Epoxy resins, only of the best quality, are used sparingly. Indium or gold seals are preferred, but if rubber O rings are used, like here, these must also be outgassed thoroughly.

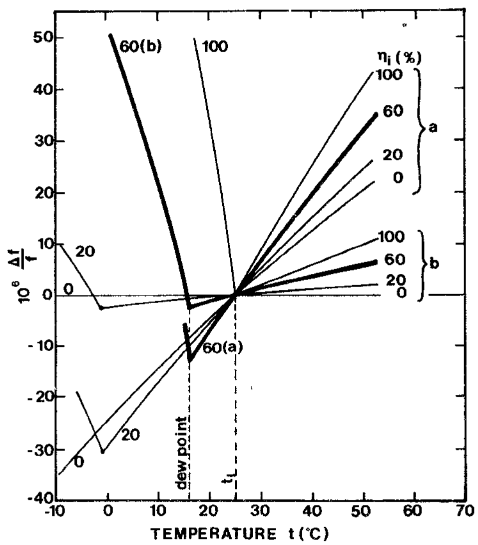

Fig. 4. Calculated variation of the resonant frequency with temperature, for a cavity filled with air of different humidity at 25°C. Curves a: unsealed cavity. Curves b: sealed cavity.

Since it is one of the sources which influence temperature stability, the diode may not be disregarded. The diode impedance changes with temperature, caused by the variation of electron and hole drift velocities, ionization coefficients, and of the junction capacitance and series resistance for a nonpunched-through diode. If, for the diode considered, we take into account the doping concentrations and profiles, known by measurements, a calculation shows that for a low-Q Iglesias-type oscillator, the final frequency shift must be in the order of -100 ppm/°C. This value has been verified experimentally. With a stabilization factor $S \approx 100$ this means a variation of -1 ppm/°C for the Kurokawa oscillator, which, of course, must be compensated for. Further, it has to be noted that some types of passivation layer on the diode chip, especially those of organic chemical nature, may cause an initial (often positive) frequency drift caused by partial thermal decomposition.

V. Measurements

The results of measurements on both types of oscillators are summarized in Table I. The circuit coupling parameters and Q values indicated were evaluated by standard bridge measuring techniques. Moreover, a check on the coupling parameters was obtained by measuring the power dissipated in the coaxial load and in the output load. In combination with a measurement of the power, generated by the diode in a coaxial Iglesias-type oscillator at the same frequency and diode impedance level, it is then possible to evaluate β_1, β_2, and η_c. Completed with a pulling measurement, yielding a value for Q_{ext}, a good agreement was obtained with the results of the cold measurements. Fig. 5 represents the measured FM noise for baseband frequencies from 10 kHz to 10 MHz (distance to carrier). For ease of comparison with data presented in [1], the measured Δf_{rms} is also transferred into a signal-to-noise ratio and noise power, corresponding to 1800-channel microwave link test

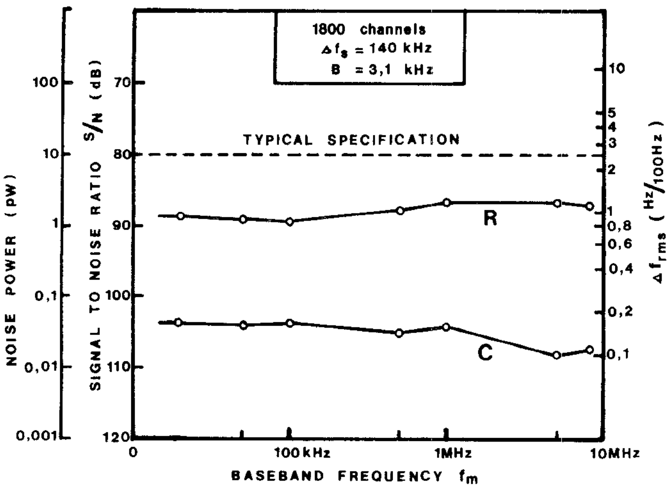

Fig. 5. Measured FM noise of the *R*- and *C*-type IMPATT oscillator as a function of the baseband frequency f_m. *R* circuit: f_0 = 7.1 GHz, P_0 = 55 mW. *C* circuit: f_0 = 6.75 GHz, P_0 = 50 mW.

TABLE I

oscillator tuning range	R-circuit 7.0 - 7.4 GHz	C-circuit 6.4 - 7.1 GHz
β_1	11	9
β_2	0.80	0.55
η_c	41%	33%
Q_o	6000	24800
$Q_{ext} = \frac{Q_o}{\beta_2}$	7500	45100
$Q_L = \frac{Q_o}{1+\beta_2}$	3330	16000
$I_{d.c.}$	40 mA	40 mA
$V_{d.c.}$	130 V	130 V
$I_{start\ osc.}$	28 mA	20 mA
P_{out}	55 mW	50 mW
pulling (VSWR=1.5)	500 kHz	100 kHz
pushing	3 kHz/mA	0.7 kHz/mA
FM noise in 100 Hz at f_m = 500 kHz $\Delta f_{r.m.s.}$	0.84 Hz	0.17 Hz
frequency stability	see Figure 6	

conditions, supposing no preemphasis and no psophometric weighting. The conversion formulas are then

$$\text{noise power} = 1.58(\Delta f_{\text{rms}})^2 \qquad \text{pW}$$

and

$$S/N = 88 - 20 \log \Delta f_{\text{rms}} \qquad \text{dB}.$$

According to (5) the calculated stabilization factors for the *R* and *C* circuits are

$$S_R = \frac{3080}{Q_{0g}} \qquad S_C = \frac{14\,700}{Q_{0g}}.$$

Their ratio is 4.8, in agreement with the measured figures for pushing and FM noise. The output FM noise Δf_{rms}, measured in a low-Q coaxial oscillator (Iglesias circuit) was 80 Hz/100 Hz. Taking into account the values, measured on both oscillators, it is concluded that the stabilization factors are $S_R \approx 100$ and $S_C \approx 500$ for the *R* and *C* circuit, respectively. This, in turn, gives $Q_{0g} \approx 30$, which is a plausible figure.

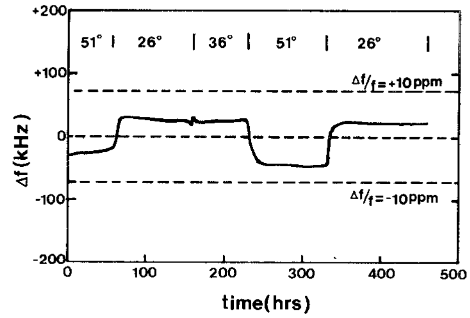

Fig. 6. Measured long-term frequency stability of *R*-type IMPATT oscillator. f = 7348 MHz + Δf, P_0 = 55 mW (25°C). Parameter: ambient temperature in degrees Celsius.

Time- and temperature-dependent frequency stability tests were performed on an *R*-type oscillator at a frequency of 7348 MHz. As can be seen from Fig. 6, temperature cycling of this oscillator between ambient temperatures of 26 and 51°C over a long period (450 h), gave a frequency shift of 70 kHz, corresponding to an average frequency stability better than −0.4 ppm/°C, with substantially no drift during intervals of constant temperature.

VI. Conclusions

1) By proper choice and design of a high-Q stabilizing circuit, it is demonstrated that, even when using a rather noisy device like the Si IMPATT diode, stable oscillators can be realized with a very low FM noise level.

2) These oscillators easily fulfill the noise specifications for use as a local oscillator in high-capacity FM microwave links, handling telephone or TV signals.

3) With respect to noise performance in the lower baseband slots and to power output, the IMPATT oscillators described are superior to usual Gunn oscillators.

4) Proper design of temperature compensation means and special measures taken to eliminate changes of the medium inside the cavity have shown the feasibility of frequency stabilities better than ±1 ppm/°C.

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