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Classification
Physics Abstracts
05.30 — 23.00 — 23.20 — 23.40 — 23.50

EXPERIMENTAL Q FACTORS OF THREE TYPES OF MICROSTRIP RESONATORS

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(*Reçu le 9 octobre 1972, révisé le 24 janvier 1973*)

Résumé. — Des mesures du coefficient de surtension Q ont été réalisées sur trois types de résonateurs microbandes à substrat d'alumine : ligne ouverte de longueur λ, anneau de longueur 3λ et disque. La comparaison des valeurs expérimentales et théoriques est satisfaisante. Pour la dernière structure, un Q_0 de 900 a été obtenu.

Abstract. — Unloaded Q factor measurements have been made of one-port integral-wavelength open line, a three-wave circular ring, and one empty disc resonator on an Al_2O_3 substrate. Theoretical and measured values are compared. Q_0 factors up to 900 have been measured in the disc resonator.

Microstrip resonators which use linear [1], [2], ring [3], [4] and disc [5], [6] structures are well known (Fig. 1). They are examined here in order to obtain the highest value of the unloaded quality factor Q_0 and to build very compact oscillators.

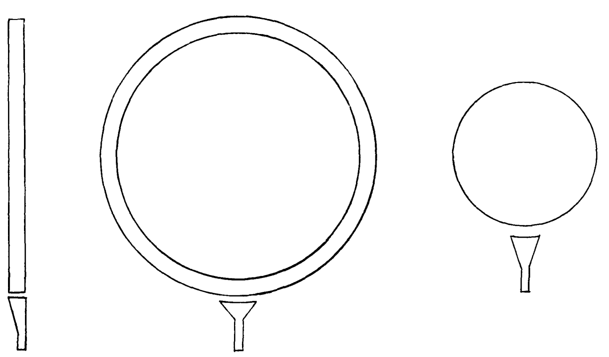

FIG. 1. — Plan view of the three types of resonators and coupling probes. Impedances : coupling probes 50 Ω, resonators : *a*) 30 Ω ; *b*) 30 or 50 Ω. Dimensions : *a*) line length $l = n\frac{\lambda}{2}$; *b*) ring mean length $l = n\lambda$; *c*) disc diameter $d = \frac{55,23}{\pi f \sqrt{\varepsilon_r}}$ cm.

Resonant structures with one port have been utilized (Fig. 1). The coupling between the 50 Ω line input and the resonator changes with the gap thickness ; we have optimized this gap for a coupling factor $\beta \simeq 1$. The probe is always realized with a tapered linear section to match the 50 Ω line to the linear resonator and to assure the best coupling between the line and the ring or disc.

The quality factors were measured by classical reflection methods [7], [8] using a type 8542 Hewlett-Packard Automatic Network Analyzer ; the reflection resonator method has been preferred to the transmission one [1], [3] for the reasons given by Pucel *et al.* [9].

Figure 2 illustrates the method used to calculate the resonator loaded Q, Q_L, and the coupling factor β, as it was proposed by Ginzton [7]. Q_L may be obtained by considering on the Smith chart the segments connecting the points $Z = 0 \pm j\,0$ and $Z = 0 \pm j\,1$, which correspond to the half power (3 dB) points. Q_L is defined, by the formula

$$Q_L = \frac{f_0}{f_1 - f_2} = \frac{f_0}{\Delta f},$$

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Article published online by EDP Sciences and available at http://dx.doi.org/10.1051/rphysap:0197300803027900

where Δf is the 3 dB bandwidth and f_0 the resonating frequency ; β is equal to r_0 or $1/r_0$ for an overcoupled or undercoupled resonator respectively.

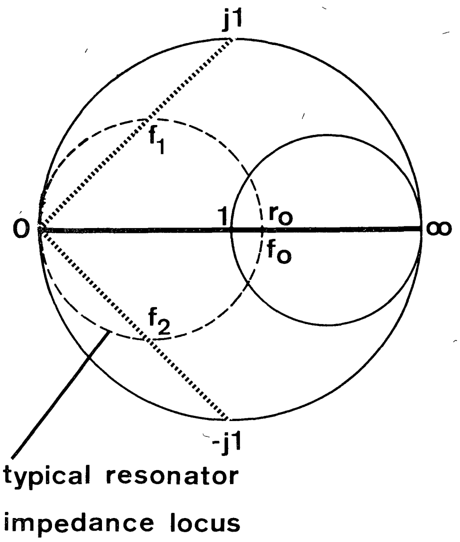

FIG. 2. — Typical resonator impedance locus versus frequency.

The locus and lines may be drawn on the Smith chart overlays for the polar display on the network analyser. Q_L may be obtained by centering f_0 on the real axis of the Smith chart and then adjusting Δf so that the end points of the sweep are on the Q_L lines drawn on the overlay : f_0 and Δf are adjusted by using the sweeper so that the resonance response is symmetrical about the real axis on the Smith chart. Knowing Q_L and β we may obtain

$$Q_0 = Q_L(1 + \beta) .$$

This method proved to be very fast and accurate for Q_0 measurements between 100 and 1 000, compared to the classical and tedious VSWR method. The measurements system allows the complete characterization of the resonator with no degradation in accuracy and needs a shorter time than a point by point measurement method.

The reflection measurement yields both magnitude and phase information which fully describes the microstrip resonator terminating the transmission line. Our reflectometer system is essentially formed by :

— a sweep oscillator, to plot the cavity impedance locus versus frequency ;

— a transducer unit with two directional couplers, for reference and test channels, which provides the capability of extending the electrical length of the reference channel so that the respective paths of the reference and test signals are equal ;

— a harmonic frequency converter in which the input signal is converted by harmonic sampling to a fixed IF frequency so that low frequency circuitry can measure amplitude and phase relationships ;

— an output display unit : polar display (Smith chart) for a fast measurement. The two frequencies f_1 and f_2 are directly read on the sweeper oscillator.

Theoretically, substrate thickness, dielectric losses, surface roughness and glaze, and plating thickness and resistance must all be accounted for in calculating microstrip attenuation, but dielectric losses are about a factor of 100 less than resistive losses in the conductor [1], [10] and they have therefore been neglected. Only the conductor's Q_0 is calculated and compared with experimental one.

For the Q_L measurements, the resonator is placed in a shielding box to avoid radiation losses. The parasitic resonances of the enclosure are avoided by using a thick film of « aquadag » on the cover plate. However, to estimate radiation losses of each resonant structure, Q_L measurements are performed with the microstrip in its box, with the cover removed.

1. **Linear resonator.** — Various open-circuit integral half-wavelength line sections (Fig. 1*a*) were realized at a resonant frequency close to 3 GHz. Their physical parameters are tabulated in table I.

The conductor Q_0 of a microstrip line resonator is given by

$$Q_0 \simeq \frac{27.3}{\lambda_0\, \alpha_0} \tag{1}$$

TABLE I

Parameter	Dimensions	Resonator Linear	Ring 1	Ring 2	Disc
Alumina thickness, h	mm	1.0	1.0	1.0	1.0
Line width, w, or disc diameter, d	mm	2.4	2.4	1.0	19.7
Line thickness, t	μm	10	10	10	10
Alumina purity		99.5 %	99.5 %	99.5 %	99.5 %
Alumina surface roughness	μm	1.0	1.0	1.0	1.0
Dielectric constant, ε_r		9.6	9.6	9.6	9.6
Characteristic impedance, Z_0	Ω	30	30	50	—
Line length, l	mm	37.2 (1)	112.5 (2)	118.2 (2)	—

(1) $l = \lambda$ at 3 GHz.
(2) $l = 3\,\lambda$ at 3 GHz.

where λ_0 is the wavelength (m) and α_0 the attenuation per unit length (dB/m) in air. α_0 is a function of microstrip physical parameters and current density distribution across the conductor width [11]:

$$\alpha_0 = \frac{Z_0 R_s}{720 \pi^2 \log_e 10} \left[\frac{1}{h} + \frac{0{,}44\, h}{w^2} + \frac{6\, h}{w^2}\left(1 - \frac{h}{w}\right)^5\right] \times \left(1 + \frac{w}{h} + \frac{1}{\pi} \log_e \frac{2\, h}{t}\right)$$

in dB/unit length, for $w/h \geqslant 1$. Z_0 is the characteristic impedance, R_s the skin resistance of the metal, h the alumina thickness, w the line width and t the line thickness. For a uniform current distribution in the conductor, α_0 would be given by the expression:

$$\alpha_0 = \frac{20}{\log_e 10} \frac{R_s}{w Z_0}.$$

Theoretical values of Q_0 for a uniform (Q_{0u}) and nonuniform (Q_{0nu}) current distribution and measured Q_0 are shown versus frequency in figure 3 for $w/h \geqslant 1$. There is a great difference between Q_{0nu} and the experimental value, but a difference of less than 10 % is obtained between Q_{0u} and the measured Q_0. The theoretical Q_{0u} is a good approximation of the actual Q_0 because the additional losses are not taken into account in the theoretical analysis.

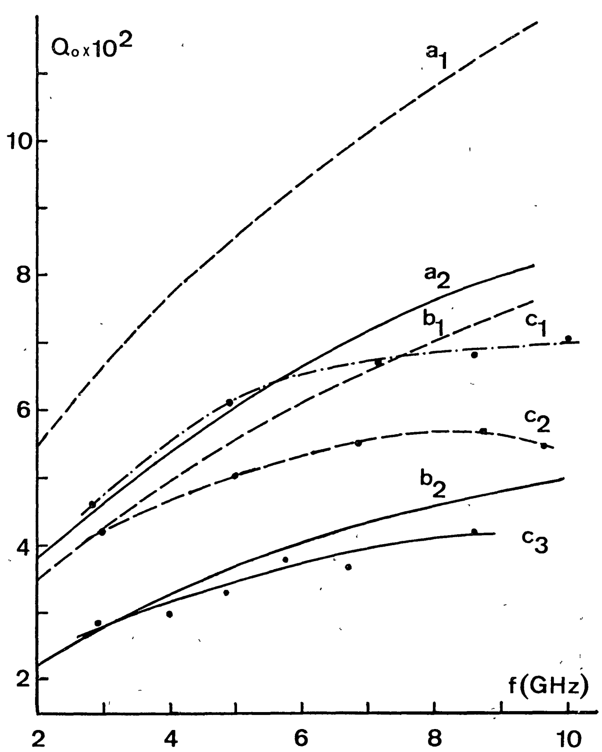

FIG. 3. — Unloaded Q for open line and ring resonators. *a*) theoretical Q_{0nu} (a_1, 30 Ω ; a_2, 50 Ω) ; *b*) theoretical Q_{0u} (b_1, 30 Ω ; b_2, 50 Ω) ; *c*) experimental Q_0 (c_1, open line ; c_2, ring 1 ; c_3, ring 2)

——— $Z_0 = 50\,\Omega$;
– – – / — · — . } $Z_0 = 30\,\Omega$.

2. **Ring resonator.** — Two different integral half-wavelength line sections were realized in circular ring structures (Fig. 1*b*) having characteristic impedance of 30 and 50 Ω respectively. An electric length $l = 3\,\lambda$ at 3 GHz was used to avoid the effects of mutual inductance. Their physical parameters are tabulated in table I.

Figure 3 shows theoretical values of Q_{0u}, Q_{0nu} and measured Q_0 versus frequency for both resonators ([1]). As for the linear resonator, a great difference between Q_{0nu} and the experimental values are obtained but there is quite a good agreement between the theoretical Q_{0u} and the measured Q_0 for the highest Z_0 (full curves b_2 and c_3 for $Z_0 = 50\,\Omega$ of Fig. 3).

The differences reported between Q_{0nu} and the measured linear or ring resonators are due to our simple theoretical model where only constant conductor losses are considered. These discrepancies can be lessened if one takes into account the surface roughnesses by adding a sizeable correction [9], and radiation losses. The theoretical Q_{0nu}, corrected as mentioned above, is a limit value.

A comparison between the 30 Ω linear and ring resonators (curves c_1 and c_2 respectively of Fig. 3) shows that the measured ring Q_0 is smaller than the measured linear resonator Q_0. This effect cannot be explained by a difference in the radiation losses because we have established experimentally that the radiation losses in the open line are up to 50 % higher than in the ring when the top of the shielding box is removed : the open end of the line radiates [12], [13], whereas the ring is a closed structure and its radiation losses are negligible. This difference could be explained by a smaller line length of the linear resonator (see Table I) ; α_0 would then smaller for the linear than for the ring resonator. We are examining this problem.

Concerning the resonance frequency of the ring resonators, we have observed, in agreement with Wolff and Knoppik [14], a dispersion between the theoretical and measured frequencies as shown by figure 1 of Wolff *et al.* [14]. Our aluminas have an $\varepsilon = 9.6$ and a curve like their « a » is obtained for ring 2 (50 Ω). For the ring 1 (30 Ω) this curve is placed between their « d » and « e ».

(1) The measured Q_0 for ring 1 can be compared with types B and C of Troughton [4] for frequencies up to 8 GHz. Troughton used $Z_0 = 25\,\Omega$, alumina surface roughness about 0.23 μm, $\varepsilon_r = 9.9$ and alumina purity lower than 99.9 %, probably 99.5 %.

3. **Disc resonator.** — The unloaded quality factor of this resonator (Fig. 1*c*) is given by [6] :

$$Q_0 = h(\pi f \mu_0 \sigma)^{1/2} = 4.8 \times 10^2 \ hF^{1/2} \qquad (2)$$

with

F = frequency in GHz = 10^9 f.
h = alumina thickness in mm.
σ = conductivity of copper = 5.8×10^7 ℧.m^{-1}.
$\mu_0 = 4\pi \times 10^{-7}$ H.m^{-1}.

Resonators with a fundamental frequency of 3 GHz for the mode $n = 1$ have been realized, and their physical parameters are tabulated in table I.

Figure 4 shows theoretical and experimentally measured Q_0 values versus frequency for the fundamental frequency (mode $n = 1$) and other modes (n = 0, 2, 3). The measured values are always smaller than the theoretical ones by about 30 %.

A comparison between figures 3 and 4 shows that, for identical alumina purity and surface polishing and the same line and alumina thickness (Table I), the experimental Q_0 are higher for disc resonator than for other geometries ; it is possible to obtain Q_0 = 900 at 7.7 GHz (n = 3) with a disc. Another interesting feature of this resonator is the possibility of matching the impedance of a diode. For this reason, many researchers put an IMPATT or GUNN diode in the center of the resonator, but it appears still more interesting to couple it to the periphery of the disc and vary the electromagnetic field configuration capacitively to obtain the maximum output power.

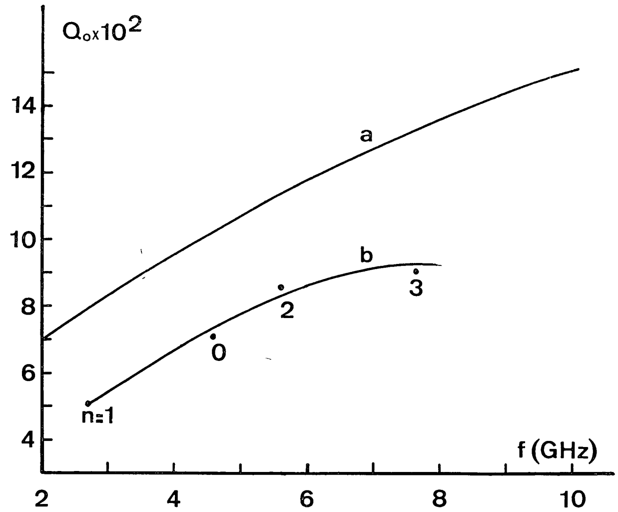

FIG. 4. — Unloaded Q for disc resonator : *a*) theoretical ; *b*) experimental for fundamental frequency (n = 1 mode) and harmonics (n = 0, 2, 3 modes).

4. **Conclusions.** — We have established that radiation losses are higher in the disc than in the linear resonator, as reported by Delinger [13], for a non-shielded configuration. This effect is not included in the analysis of Watkins [6]. The smallest radiation losses are obtained with a ring structure.

Finally, in the open line the calculated and experimental Q_{0u} are relatively close to each other, and there is quite good agreement with Q_{0u} for the highest Z_0 ring. The theoretical Q_{0nu} corrected by additive losses due to surface roughness and radiation is an upper limit.

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