Design of Microwave Oscillators and Filters Using Transmission-Mode Dielectric Resonators Coupled to Microstrip Lines

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Abstract —**A more detailed model for the transmission-mode dielectric resonator coupled between microstrip lines is given. Novel design approaches for parallel feedback oscillators and bandpass filters are discussed. For oscillators, the design mainly takes into account zero phase shift loop considerations, as in the classical low-frequency approach. Oscillators of this type may offer low phase noise. For filters, the spatial separation between dielectric resonators favors multipole designs. Using the same microstrip layout, different shapes and bandwidths may be obtained by simple tuning.**

I. Introduction

DIELECTRIC RESONATORS (DR's) have brought significant improvement in the design of microwave oscillators and filters [1]. Compactness, light weight, temperature stability, and relative low costs are easily achieved.

For oscillators, active two-port devices (GaAs FET's and bipolar microwave transistors) are usually preferred for their efficiency. Microstrip technology is also preferred because integration is easily obtained.

For filters, integration is also possible. Designs using microstrip input and output lines were shown to be related [2], [3].

For oscillators, several basic configurations are possible. The DR may be placed either at the output [4] or at the input [5] of the active device. However, broader tuning bandwidths are supposed to be obtained if the DR works as a parallel feedback element. Ishiara *et al.* [6] have presented this type of design. More recently, Khanna [7] also used the DR as a parallel feedback element. Emphasis was placed on obtaining maximum values of the small-signal output reflection coefficient. Neither of these two papers mentioned above, dealt with controlling the total phase shift around the feedback loop.

In what concerns bandpass filters, the papers usually relate a direct electromagnetic coupling from side-by-side DR's [1], [3], [8]. In practice, the final distances between adjacent resonators must be somewhat readjusted. For multipole filters —a large number of DR's— this readjustment becomes a nuisance as it disturbs the whole DR chain. Alternatively, Guillon *et al.* [9] suggest using a simple (two-pole) bandpass filter cascaded with two other bandstop filters. The performance of a higher order filter is then achieved.

In this paper, a more complete model for a transmission-mode DR, coupled between microstrip lines, will be presented. For the design of oscillators, it will be possible to prescribe the magnitude and the phase of the feedback factor. The design may be carried out aiming at obtaining a highly efficient oscillator or a broad-band one.

For the design of bandpass filters, it will be possible to couple each DR to an individual microstrip section instead of a side-by-side DR arrangement. Spatial separation between the several DR's may lead to an easier multipole design. No mutual interference between the DR's will be experienced. Each one is easily adjusted either with respect to its own coupling factor or tuned with respect to its own resonant frequency.

II. Model Proposed

A model for the DR coupled between two microstrip lines and functioning at the transmission-mode has been previously presented [10]. This latter, however, is limited to a particular situation; the distance θ from the DR to the microstrip line edges is a quarter of a wavelength, as shown in Fig. 1(a). Therefore, strictly speaking, there is another possible configuration for a DR coupled between microstrip lines. This configuration is the one depicted in Fig. 1(b). There are some electrical differences between the two above mentioned arrangements. Complete models for these two circuits will be given now for any value of θ.

From electromagnetic considerations and from measured results, the authors are proposing that the two above circuits may be modeled as given in Fig. 2(a) and (b), respectively. From these models, it is possible to distinguish an initial difference between the two arrangements: a phase inversion exists as far as transmission is concerned. Another difference is that the resonant

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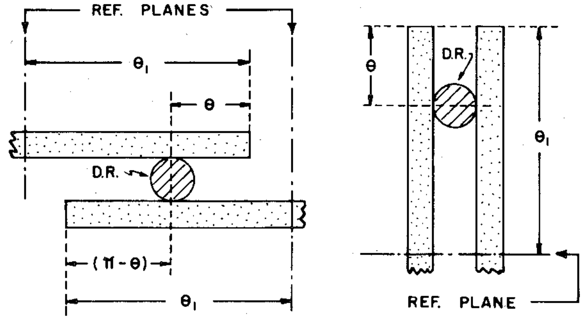
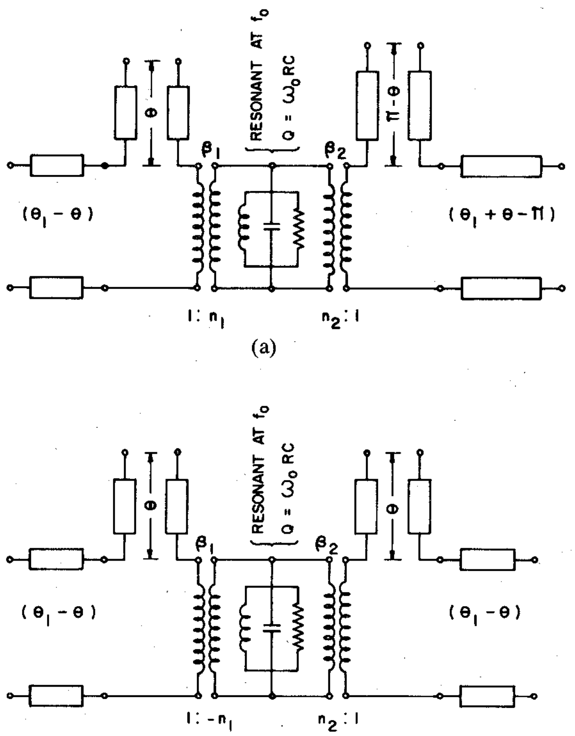


(a) (b)

Fig. 1. Two distinct configurations for transmission-mode DR coupled between microstrip lines. (a) Input and output lines from opposite directions. (b) Input and output lines from same direction.



(b)

Fig. 2. Equivalent circuits using open series stubs. (a) Model for Fig. 1(a): noninverting arrangement. (b) Model for Fig. 1(b): phase inverting arrangement. All transmission lines present a characteristic impedance equal to Z_0, the system reference.

frequency of the circuit of Fig. 1(a) is, for all θ, the resonant frequency of the DR itself. For the arrangement of Fig. 1(b), the resonant frequency is very near, but not equal to, that of the DR. Furthermore, this last frequency depends upon θ, the relative position of the DR. However, at this point, this last circuit will be constantly retuned to keep the same resonant frequency.

When the amplitude response is measured with respect to θ, the results obtained, for both circuits, are those depicted in Fig. 3; a continuous variation from a transmission zero up to a maximum value, occurring at $\theta = \pi/2$. This maximum value depends upon the coupling factors, β_1 and β_2, between the DR and the microstrip lines.

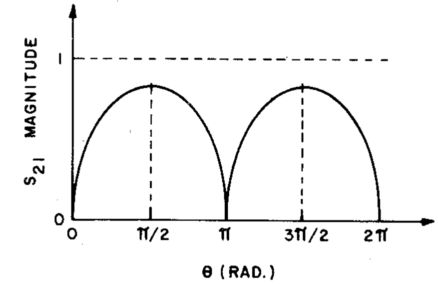


Fig. 3. Transmission coefficient magnitude as a function of the DR position.

When the phase response is measured, and the resonant frequency is kept constant, it is observed that both circuits provide a constant phase for all θ.

For the symmetrical case $\beta_1 = \beta_2 = \beta$, the following equations describe the transmission and reflection coefficients of both arrangements. For the arrangement of Fig. 1(a)

$$(1/S_{21}) = -\{[1+(1/a)]+j(\delta/a)\}\cdot p$$
$$S_{11} = (1/p) + S_{21}.$$

For the arrangement of Fig. 1(b)

$$(1/S_{21}) = \{[1+(1/a)]+j[(\delta/a)+\cot\theta]\}\cdot p$$
$$S_{11} = (1/p) - S_{21}$$

where $\delta = Q[(f^2/f_0^2)-1]$; $a = 2\beta\sin^2\theta$, the phase factor $p = \cos 2\theta_1 + j\sin 2\theta_1$; θ_1 is related to the reference plane distances as shown in Fig. 1, f_0 is the resonant frequency of the DR itself, while Q is the unloaded DR quality factor in the MIC environment.

III. Oscillator Design

By using the above-described model, an oscillator design is easily accomplished. It is a matter of providing access lines, from input and output of the active device, in a way that the total phase shift is an integer multiple of 2π. The active device transmission coefficient S_{21}^a is included in this computation, together with that of the specific DR arrangement. The magnitude loop gain is also easily made slightly greater than the unity by using a suitable value of θ. It is interesting to remember that further small θ adjustments will not interfere with the loop phase shift, rather only with the gain value. For feedback phase loop computation, the small-signal value of S_{21}^a may be used as a first approximation.

Next, a fine tuning of θ_1, concerning the length of the microstrip lines, may provide optimum large-signal phase loop conditions. Hence, the resonant circuit may work exactly at its resonant frequency. As in resonance, the slope of phase variation, with respect to the frequency, is maximum, and the phase noise will be minimum [11].

Two MIC 4-GHz oscillators, using the bipolar NEC 56708D transistors, are offered as examples. The substrate chosen presents $\epsilon_r = 10.5$ and a height of 0.63 mm. Both DR's are equal and have $\epsilon_r = 37.7$ with a diameter of 15.0 and height of 6.0 mm. The feedback factor was used as a

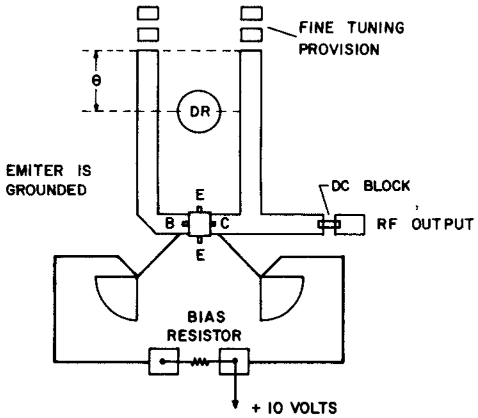


Fig. 4. Typical layout of a feedback DR stabilized oscillator.

TABLE I
SUMMARY OF RESULTS OBTAINED WITH 4-GHz PARALLEL FEEDBACK OSCILLATORS USING THE FEEDBACK FACTOR AS A PARAMETER

Nominal feedback factor	Power output	Tunability mechan.	Efficiency	Phase noise at 10-KHz offset
−5 dB	+19 dBm	80 MHz	27%	−100 dBc/Hz
−2 dB	+13 dBm	300 MHz	20%	−95 dBc/Hz

parameter for these two oscillators. The first one uses a relatively loose feedback factor, while the second a tighter one. Excepting the feedback factor, both layouts are therefore similar and are shown in Fig. 4. In the first oscillator, the separation between microstrip of DR edges is 2 mm. In the second one, this separation was reduced to be about 0.5 mm. Results of measures obtained are those presented in Table I.

If just the right amount of power is routed through the feedback loop, a relatively high efficiency is achieved, as described in the first case. By carefully tuning phase conditions, the measured phase noise was −100 dBc/Hz at 10-KHz offset of the carrier. This last figure is a very good one and comparable with recent results mentioned in the literature [12], [13], when no special features for noise reduction were used.

By using a tighter feedback coupling factor, as it was done in the second oscillator, it is possible to increase the tunability range. In this example, a 7.5-percent relative bandwidth was easily obtained. In other cases, by using progressively tighter feedback factors, bandwidths beyond 10 percent were reached. However, efficiency will suffer progressively as will the phase noise.

IV. FILTER DESIGN

Within the scope of the above-described model, and keeping θ as a parameter, it is easy to characterize the DR with respect to its slope parameter [13]. The filter design is then easily accomplished by quite classical methods. Simple $\lambda/4$, or $3\lambda/4$, lines are used as impedance inverters.

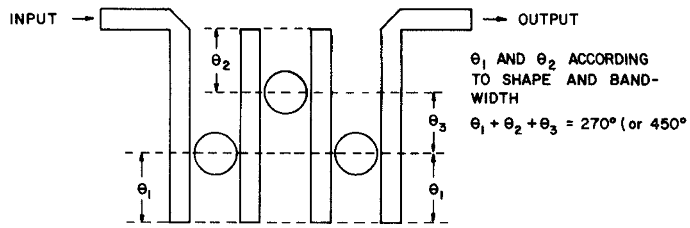


Fig. 5. Typical layout of a three-element bandpass filter using transmission-mode dielectric resonators coupled between the microstrip lines.

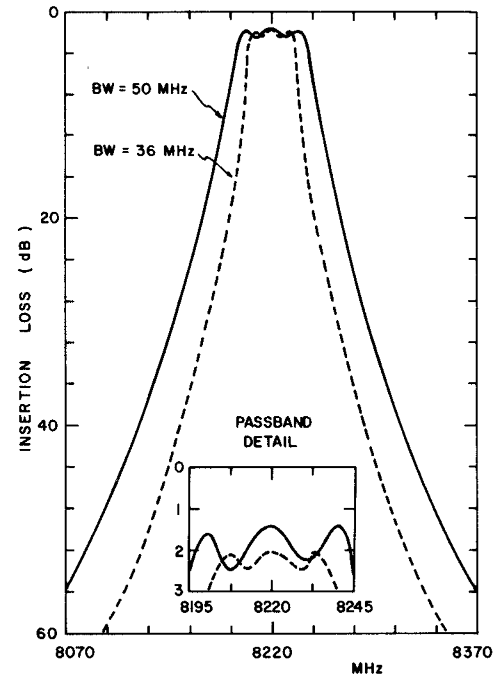


Fig. 6. Measured performance of two filters obtained when the DR's are repositioned over the same layout. Both filters are centered at 8220 MHz. Dashed line: 36-MHz bandwidth, 0.5-dB Chebyshev. Solid line: 50-MHz bandwidth, 1.0-dB Chebyshev.

Here, as a first demonstration, microstrip step discontinuities were avoided. Only 50-Ω lines were used. One of the possible aspects of a three-element filter layout is presented in Fig. 5. If the central passband frequency is kept constant, then small variations of the θ's involved may modify somewhat the filter's shaping and bandwidth. To show this design's versatility, two practical results are furnished; both are centered at 8200 MHz. Substrate was chosen with $\epsilon_r = 2.2$ and height of 0.8 mm; DR's have $\epsilon_r = 36.7$ and dimensions of 6.2 and 4.3 mm for diameter and height, respectively.

The first result is from a 36 MHz wide, 0.5 dB, Chebyshev filter. The insertion loss was measured, as shown in Fig. 6 (dashed line) providing more than 60 dB of rejection at an 150-MHz offset. Over the same hardware, the DR's are repositioned. An 50 MHz, 1 dB Chebyshev is then obtained. The result is presented in the same figure as a solid line.

As the filters reach broader bandwidths, it was observed that the insertion loss decreases. From several trial filters, at different frequencies, it was possible to observe that the practical workable relative bandwidth range is from 2 to 0.3 percent. Typical insertion losses are about 0.2 to 0.7 dB per resonant element, respectively.

V. Conclusions

A more complete model for the microstrip-coupled transmission-mode DR was presented. The coupling coefficient is easily changed. Parallel feedback oscillator design may aim at efficiency and low phase noise or broad tuning range. No special microelectronics facilities are needed for obtaining a microwave oscillator from encapsulated transistors.

Filter design features flexibility of modifying shape and bandwidth while keeping the same hardware. Spatial separation between the DR's favors multipole design. Filter tuning is easily accomplished.

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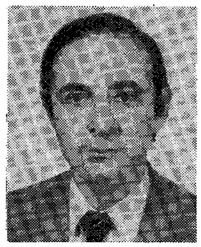

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