DIELECTRIC RESONATOR OSCILLATORS THEORY AND STATE–OF–THE–ART*

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INTRODUCTION

With the advent of high–Q temperature–stable dielectric material, the transistor dielectric resonator oscillator (DRO) is fast becoming the most desirable choice in a vast number of fixed–frequency microwave signal source applications. A DRO clearly fills the gap between a free–running oscillator and a synthesized source, and represents a good compromise of cost, size and performance compared to alternative signal sources such as high–Q cavity oscillators, microstrip oscillators or multiplied crystal oscillators. Dielectric resonators, due to their excellent integrability in MIC circuits, can directly be used as the frequency–determining element for realizing a stable microwave frequency source. The primary characteristics of the ceramic material to be used for dielectric resonators are:

- The quality factor, Q, which is approximately equal to the inverse of the loss tangent.

- The temperature coefficient of the resonant frequency, τ_f which includes the combined effects of the temperature coefficient of the dielectric constant and the thermal expansion of the dielectric resonator and the shielding package.

- The dielectric constant, ϵ_r.

The Q, τ_f, and ϵ_r values required for various applications differ and, in general, satisfactory oscillator operation under most conditions can be achieved by choosing an appropriate material composition. Until several years ago, the lack of suitable materials (i.e., materials possessing acceptable combinations of Q, τ_f, and ϵ_r) severely limited dielectric resonator applications. Materials such as rutile–phase TiO_2, which has an unloaded Q of about 10000 at 4 GHz and ϵ_r of 100, were most often used. However, TiO_2 has a τ_f value of 400 ppm/°C which makes it impractical for most applications.

The development of temperature–stable dielectric resonators dates back about a decade. A number of material compositions have been explored in attempts to develop suitable dielectric materials, including ceramic mixtures containing TiO_2, various Titanates and Zirconates, glass, ceramic systems and alumina–based ceramics. At present, several ceramic compositions have been developed offering excellent dielectric properties. Complex perouskite compounds with the general formula $A(B'_{1/3}B''_{2/3})O_3$

(where $A = Ba,\ Sr$; $B' = Zn,\ Mg,\ Co,\ Ni$; $B'' = Ta,\ Nb$) have proved to possess acceptable properties for dielectric resonators. These compounds have dielectric constants between 20 and 40, a high quality factor, some greater than 10000 at 10 GHz, and a temperature coefficient which is variable through modification of the composition. Table 1 compares the important properties of different materials developed commercially[1].

It is impossible to say that any of the dielectric compositions shown in Table 1 is "better" than any other, since other factors, such as the ease of ceramic processing and the ability of a manufacturer to hold tolerances on the dielectric properties must also be considered. Performance limitations, if any, of the lower dielectric constant materials remain to be determined, since most component work reported thus far has used dielectric resonators with ϵ_r in the range of 37–100. The lower dielectric constant material is likely to be more sensitive to shielding, due to the increase in fields outside the resonator.

RESONANT FREQUENCY

A dielectric resonator, made of low–loss, high–permittivity ceramic material, resonates in various modes at frequencies determined both by its dimensions and its surroundings[1]. Although the geometrical form of a dielectric resonator is extremely simple, an exact solution of the Maxwell equations is considerably more difficult than for the hollow metallic cavity. For this reason, the exact resonant frequency of a particular resonant mode, such as the most commonly used $TE_{01\delta}$, can only be computed by rigorous numerical procedures.

A number of mathematical techniques which can predict resonant frequency to an accuracy of $\pm 1\%$ for specific configurations appear in the literature. Unfortunately, these methods call for the use of high–powered computers. Kajfez[2] has presented an approximate solution of the equations for both the case of an isolated dielectric resonator and for a resonator coupled into a MIC microstrip circuit. This method is typically accurate to $\pm 2\%$. Practical dielectric resonators available today cover the frequency range of 2 to 100 GHz. However, with the recent development of conveniently–sized coaxial tubular dielectric resonators, the usable frequency range has been extended to 500 MHz. The lower frequency limit is imposed by the large resulting dimensions of the resonator, while the upper frequency is limited by the reduced Q of small resonators, as well as by resonator dimensions that become too small to effectively couple to a transmission line.

To a first approximation, a dielectric resonator is the dual of a metallic cavity. The radiation losses of the dielectric resonators with commonly–available permittivities, however, are generally much greater than the energy losses in the metallic cavities, which makes proper shielding of the dielectric resonator a necessity. Dimensions of a dielectric resonator are also considerably smaller than those of an empty metallic cavity resonant at the same frequency – by a factor of approximately $\sqrt{\epsilon_r}$. If ϵ_r is high, the electric and magnetic fields are confined in the region near the resonator, which results in small radiation losses. The unloaded quality factor is thus limited by the losses in the dielectric resonator. The shape of a dielectric resonator is usually a solid cylinder, but tubular, spherical and parallelopiped shapes are also used.

In the $TE_{01\delta}$ mode, the most commonly used mode, magnetic field lines are contained in the meridian plane while the electric field lines are concentric circles around z–axis as shown in Fig. 1. For a distant observer, this mode appears as a magnetic dipole, and for this reason it is sometimes referred–to as the "magnetic dipole mode." When the relative dielectric constant is around 40, more than 95% of

the stored electric energy, and more than 60% of the stored magnetic energy is located within the cylinder. The remaining energy is distributed in the air around the resonator, decaying rapidly with distance away from the resonator surface.

COUPLING TO THE RESONATOR IS CRITICAL

To effectively use dielectric resonators in microwave circuits, it is necessary to accurately understand the coupling between the resonator and transmission lines. The $TE_{01\delta}$ mode of the cylindrical resonator can be easily coupled to a microstrip line, fin line, magnetic loop, or to a metallic or dielectric waveguide[1]. Figure 2 shows the magnetic coupling between a dielectric resonator and microstrip line. The resonator is placed on top of the microstrip substrate, with the lateral distance between the resonator and the microstrip conductor primarily determining the amount of coupling between the resonator and transmission line[3]. The degree of coupling has a direct effect upon output power, frequency stability, and harmonic content, as well as resonant frequency of a dielectric resonator oscillator. Proper metallic shielding, required to minimize the radiation losses (hence to increase Q), also affects the resonant frequency of the $TE_{01\delta}$ mode. Figure 3 & 4 show the equivalent circuit of the dielectric resonator coupled to a microstrip line.

ACTIVE DEVICES FOR DRO'S

Transistors are most commonly used to build today's DROs up to 40 GHz. Transistor DROs can be realized using either silicon bipolar or GaAs FET devices. The maximum oscillation frequency for silicon bipolar transistor oscillators is lower than that of the GaAs FET oscillators. Reasonable power outputs have been obtained up to 40 GHz and 20 GHz using GaAs FET and silicon bipolar devices respectively. Figure 5 shows the present power and frequency coverage capabilities of the dielectric resonator oscillators. The silicon bipolar transistor, however, offers lower phase noise close to the carrier and faster frequency settling characteristics compared to a GaAs FET oscillator. Other devices usable in DROs are Gunn and Impatt diodes. Gunn oscillators offer lower AM and FM noise characteristics at the cost of poor DC to RF efficiency (<1%) and reliability problems over the wide range of temperature operation. Impatts have poor noise performance, but offer medium powers at reasonable efficiencies (>30%). Transistors, on the other hand, offer medium noise with medium efficiencies (>20%). Both Gunn and Impatt diode sources are commonly used at millimeter frequencies (30–100 GHz).

DRO CIRCUITS

There are two means of incorporating a dielectric resonator in a MIC oscillator: as a passive stabilization element (stabilized DRO) or as a circuit element in a frequency–determining network (stable DRO).

A stabilized DRO is an oscillator which uses a dielectric resonator in the output plane of the circuit to stabilize an otherwise free–running oscillator (Fig. 6). This approach has several disadvantages, including a tendency toward mode jumping, frequency hysteresis problems, higher insertion loss due to the resonator being coupled to the output circuitry and increased output power variation.

The stable DRO configuration, which uses the dielectric resonator as a feedback/frequency determining element, is the most commonly used. It provides greater efficiency, simpler construction and more resistance to mode jumping and hysteresis effects than the stabilized DRO.

To realize a stable DRO, the resonator may be used as either a series or parallel feedback element in the frequency determining circuit. Figure 7 shows two common configurations of each type.

Series feedback configurations are based on the ability of the active device to produce a negative resistance (reflection coefficient greater than 1) at at least one of the three terminals, in the frequency range of interest[1][10]. The small signal oscillation conditions in this case are reduced to:

$$\left|S'_{11}\right| \bullet \left|\Gamma_1\right| > 1 \qquad \text{and}$$

$$\angle S'_{11} + \angle \Gamma_1 = 2\pi n \qquad n = 0, 1, 2, \ldots$$

where S'_{11} and Γ_1 are the reflection coefficients of the transistor and the resonator at any plane between the device and the resonator (Fig. 7a). Since Γ_1 is always less than 1, this condition implies that $\left|S'_{11}\right|$ looking into the device should be greater than 1. A distributed capacitance in the source for the configuration of Fig. 7a and inductance in the gate for the configuration shown in Fig. 7b is commonly required to generate a high value of $\left|S_{11}\right|$ (> 1). Position of the dielectric resonator with respect to the device is now determined to satisfy the oscillation condition completely.

The parallel feedback configuration is based on the use of the forward gain of a device (transmission coefficient greater than 1). In such a design, a dielectric resonator is used as a bandpass filter, and connected across the two terminals of an active device possessing transmission gain greater than the insertion loss of the dielectric resonator. To oscillate, the electrical line length between the device input and output must provide a phase shift around the feedback loop equal to an integer multiple of 2π radians at the oscillation frequency.

An advantage of the series feedback design is the relative ease of coupling to a single line, compared to the parallel circuit's requirement for coupling to two lines. In addition, the two coupling coefficients in the parallel case are not independent, increasing the difficulty of alignment. With the parallel feedback circuit, however, the use of a high–gain amplifier can allow significant decoupling of the resonator from the microstrip lines, resulting in a higher loaded Q factor with associated reduction in phase noise.

ELECTRICAL PERFORMANCE

The transistor DROs are available spanning the frequency range of 3 GHz to 40 GHz with power outputs ranging to greater than +23 dBm at X–band (Fig. 5). As noted earlier, the oscillator can use either a silicon bipolar or GaAs FET device, each device having associated tradeoffs in performance, and the oscillator can be followed by one or more buffer amplifier stages as required to meet power output specifications. The center frequency is usually specified in MHz with an associated frequency accuracy (temperature, pulling and pushing dependant) specification. The DRO frequency can be fixed, mechanically or electrically tunable over a narrow band. Important aspects of the salient characteristics of a DRO are discussed next.

TEMPERATURE STABILITY

The principal cause of DRO frequency drift with temperature is the phase deviation between the resonant circuit and the active circuit including device, feedback and output circuitry. Using the oscillation condition in the reflection coefficient form,[4] it can be easily proved that the temperature coefficient of DRO frequency is a function of the following parameters:

- The temperature coefficient, τ_f, of the dielectric resonator placed in a shielded MIC configuration.

- The unloaded Q of the dielectric resonator.

- The coupling coefficient of the dielectric resonator with the microstrip line.

- The temperature coefficient, $\tau_{\rho\eta}$, of the device (transistor) input reflection coefficient phase that is known to decrease linearly with temperature.

In order to achieve a temperature–compensated DRO, a resonator with a temperature coefficient of +1 to +4 ppm/°C is generally required to offset the negative temperature coefficient of the device phase temperature coefficient. Figure 8 shows some of the typical frequency drift curves over temperature for free–running DROs.

With present technology it is now possible to repeatably produce free–running DROs with frequency drift of less than ±100 ppm over the -55 to +85°C military temperature range, at frequencies up to 18 GHz. However, as shown in Fig. 8, results reported in the R & D environment are significantly better (±10 ppm over temperature)[4].

FREQUENCY TEMPERATURE COMPENSATION

Certain system applications require greater temperature stabilities than are achievable using free running DROs, even with temperature compensation. A number of techniques are used to improve the temperature stability of a DRO.

The Digitally Compensated DRO (DC–DRO) uses a sensor mounted in the oscillator to detect temperature changes. The output of the sensor goes through an A/D converter to produce a digital word corresponding to the particular temperature. EPROMs, programmed with the temperature characteristics of the DRO and a correction look-up table, drive a D/A converter to provide the correction signal, which is applied to the varactor of an electronically–tuned DRO. Using this technique frequency stability of ±15 ppm can be obtained over temperature[5].

The Analog–Compensated DRO (AC–DRO) uses an analog compensator circuit in conjunction with a temperature sensor to achieve up to ±20 ppm frequency stability. In analog compensation the individual oscillator is tested to produce a custom tuning voltage vs. temperature curve required to maintain a constant frequency. The compensation circuit is then aligned to fit the curve of the specific oscillator[5].

Ovenization may also be used to enhance the temperature stability of a DRO. To achieve temperature stability, the oscillator package is inserted in a temperature–stabilized oven. Using a heater element, a quick–response thermistor and associated control circuitry, the package temperature can be maintained

within ±5°C at 5 to 10 degrees above the maximum ambient temperature. A total frequency stability of better than ±10 ppm can be obtained using this approach.

Ovenized DROs offer lower phase noise than analog—or digitally—compensated DROs because the oscillator does not need to incorporate electrical tuning circuitry. Analog- and digitally–compensated DROs, however, are smaller in size and do not need the substantial amount of heater power required by the ovenized DRO.

LOCKED DRO'S

Phase Locked DROs (PL–DRO) and Injection Locked DROs (IL–DRO) are used when the requisite frequency stability and phase noise cannot be achieved using stabilization techniques. A PL–DRO or IL–DRO approach also becomes necessary when multiple oscillators are required to be phase- or frequency–coherent or both. For locked systems, a highly stable crystal–controlled signal source operating at HF or VHF is used as a reference oscillator.

In injection (frequency) locking, a VHF power amplifier driving a step–recovery diode is used to generate a wideband harmonic comb, which includes the required locking frequency. A bandpass filter is used to select the desired harmonic, and a free–running DRO is locked to the harmonic through the circuit shown in Fig. 9a[6]. The main requirement in this case is to make sure that the DRO frequency drift under all operating conditions is less than the injection locking bandwidth, Df. This bandwidth is a function of the injection power, oscillator output power, and external Q.

Injection locking is simpler and less expensive than phase locking, but the RF output is more likely to contain spurious signals at the harmonics of the reference oscillator frequency.

Figure 9b shows a typical phase–locked DRO circuit. A DC–coupled sampler/phase detector is used to mix the desired harmonic of the amplified crystal oscillator with the incoming frequency from the DRO. If the difference frequency is small enough, the loop will be driven towards a point where the difference frequency out of the sampler becomes zero. The loop then drives the DRO towards a zero phase error condition.

A search mechanism is generally included in the system so that the loop will be forced to tune through a stable lock point if the initial difference frequency is too large for capture to occur. An AC–coupled phase–locked DRO circuit is used when the output frequency of the DRO is not harmonically related to the reference oscillator frequency.

PHASE NOISE

Phase noise, related to short–term frequency stability, is characterized by variations in the output frequency which appear, in the frequency domain, as FM energy around the carrier frequency. This is shown graphically in Fig. 10. Phase noise is specified in dBc/Hz measured at specified offsets from the carrier frequency: typical offsets are 10 and 100 kHz. In a DRO, phase noise is primarily dependent on the following factors:

1. The low–frequency noise sources inherent in the active device,

2. The upconversion factor, a measure of the efficiency in the conversion of the low frequency noise to the phase noise of the microwave oscillator,

3. The loaded Q factor of the dielectric resonator,

4. The output power and external Q of the oscillator.

Optimization of phase noise performance calls for the use of a high–Q dielectric resonator and a low–noise device combined with specific design considerations in the oscillator circuit.

It has been proven that the low–frequency noise in a GaAs FET is inversely proportional to the gate length and width of the device. Both biasing conditions and the processing of the GaAs FET also play vital roles in achieving low noise oscillators. Figure 11 compares the phase noise performance of a number of different oscillators.

As noted earlier, silicon bipolar transistors are known to produce lower levels of low–frequency noise compared to GaAs FETs. Silicon bipolar DROs typically offer 6 to 10 dB improvement in the phase noise close to the carrier (up to at least 100 kHz off the carrier) compared to FET versions. Figure 11 compares the phase noise of a bipolar and GaAs FET DRO at different frequencies. Fortunately, bipolar transistors are now available for use in fundamental output oscillator circuits at up to Ku band.

Some of the newer design techniques that can be used to further reduce phase noise include:

- Low frequency feedback, using a parallel feedback circuit designed at low frequencies (up to 1 MHz) to reduce upconversion of the low frequency noise. Phase noise improvement of up to 20 dB has been reported using this technique[7]. This method is sometimes referred to as bias feedback.

- Noise degeneration, using the same dielectric resonator both as the frequency–determining element of the oscillator and the dispersive element of a frequency discriminator. The DC output of the discriminator is applied to the frequency control port of the DRO. This technique has been reported to achieve phase noise as low as -120 dBc at 10 kHz from the carrier at 10 GHz[8].

MECHANICAL TUNING

The frequency of oscillation of the dielectric resonator is dependant on a number of factors, not the least of which is its proximity to the ground plane. To take advantage of this, a tuning screw can be installed in the top cover directly above the resonator which, by reducing the distance between the resonator and the apparent ground plane, will provide for a certain amount of change in the resonant frequency (Fig. 12).

The reason for such behavior can be found in the cavity perturbation theory. Namely, when a metal wall of a resonant cavity is moved inward, the resonant frequency will decrease if the stored energy is predominantly in the electric field. Otherwise, when the stored energy close to the walls is mostly magnetic, as is the case for the shielded $TE_{01\delta}$ dielectric resonator, the resonant frequency will increase when the wall moves inward.

Current designs allow for up to 0.2% tuning range without significant degradation of other performance parameters. A properly–designed mechanical tuning option will provide a maximum of tuning range while still maintaining hermeticity and reliability, and will not appreciably affect the resonator Q factor (apparent as a degradation in noise and power performance) or temperature stability.

ELECTRONIC TUNING

Some applications, such as FMCW radar sources, narrowband–modulated communication systems or PLL systems, need electronic tuning bandwidths on the order of 0.1% to 1%. These applications require sources with low phase noise, high tuning speed and low tuning power. Electronically tunable DROs (ET–DRO) can now meet the requirements for many such applications.

ET–DROs are also commonly used for analog or digital temperature compensation of the oscillator. This application requires that the frequency tuning range of the DRO exceeds the frequency drift of the oscillator under any combination of operating conditions (temperature, load and bias variations). Various means are used to electrically tune the DRO[1], including ferrite tuning, optical tuning, and the more popular varactor and bias tuning.

Varactor tuning (a typical scheme is shown in Fig. 13) can provide up to 1% frequency adjustment. To permit varactor tuning, the dielectric resonator is coupled to another microstrip line connected to a varactor, resulting in mutually–coupled resonant circuits. The bias–voltage–dependant capacitance of the varactor varies the resonant frequency of the low–Q resonant circuit with the tuning voltage.

The amount of frequency tuning range can be controlled by varying the coupling between the low–Q microstripline/varactor circuit and the dielectric resonator circuit. Tighter coupling permits greater tuning range, however the attendant degradation in the Q factor manifests itself primarily as an increase in phase noise. Varactor tuning is by far the most common means of incorporating electronic tuning.

Bias voltage tuning takes advantage of the frequency sensitivity to changes in the supply voltage of the oscillating device. By not using an internal voltage regulator, the oscillator can be designed to provide the necessary tuning range by varying the bias voltage, typically within 0.1% of the center frequency. This is sufficient frequency variation to compensate for the frequency drift of the oscillator over load and temperature variations, as well as the long–term drift due to component aging.

Better phase noise performance can be achieved with the bias–tuned DRO than the varactor tuned DRO. The latter requires the dielectric resonator to be simultaneously coupled to two microstrip lines, thus lowering the loaded quality factor of the resonator[9]. However, as the output power is often a function of the supply voltage, care must be exercised to maintain suitable output power variation characteristics for the bias–tuned DRO.

MECHANICAL CONSIDERATIONS

The minimum size of a practical DRO is primarily limited by the cavity required by the particular resonator. Normal design procedure calls for a separation of more than one resonator diameter between the resonator and its surrounding walls in order to properly excite the $TE_{01\delta}$ resonant mode. Also, the separation between the resonator and the housing lid should be more than one resonator thickness to

minimize the effects of the lid on the resonator performance. Too-small a cavity can prevent oscillation in the preferred mode, as well as causing Q-factor degradation[3].

Hermeticity is an important consideration when defining DRO requirements. Oscillators which are backfilled with a dry inert gas and then welded closed, have less of a tendency to be susceptible to degradation due to long-term environmental effects. Hermetically-sealed oscillators are, of course, much more readily incorporated in a military system with the attendant performance and qualification requirements. Hermetic sealing is normally characterized by the helium leak rate of the DRO enclosure after it has seen a saturated helium environment. Typical leak rates of hermetically-sealed DROs are better than 10^{-7}cc(He)/second.

When a mechanical tuner is required, tuner construction, expected lifetime, over-travel protection and ease of operation are also important considerations. A properly-designed mechanical tuner should provide the desired tuning range, be simple to adjust, have an acceptable number of operations, incorporate stops to limit travel and still maintain case integrity and hermeticity.

LIMITATIONS OF TRANSISTOR DROS

Free-running DROs presently do not have the low phase noise and temperature stabilities required for certain high-performance applications. This limitation necessitates the use of phase locking, injection locking, ovenizing or analog or digital compensation circuits when necessary. Another important limitation of the DRO is the phase noise degradation under vibration due to the variation of the distance between the resonator and the outer shield. Significant deterioration of the phase noise at offsets up to the highest frequency of vibration can be expected under vibration. Ruggedization of the oscillator, injection locking or phase locking are typically used to minimize these effects.

FUTURE TRENDS

The dielectric resonator oscillator technology is developing exceedingly rapidly. Emphasis will continue to develop higher performing DROs. The following aspects are likely to be dealt with in the near future:

- Lower phase noise and higher temperature stability oscillators.

- Reduction in cost and size

- Wider mechanical and electrical tuning bandwidths

- Improvement in DRO performance under vibration

- Development of new materials for dielectric resonators for linear temperature coefficient and higher quality factor.

- Use of higher-order modes in resonators

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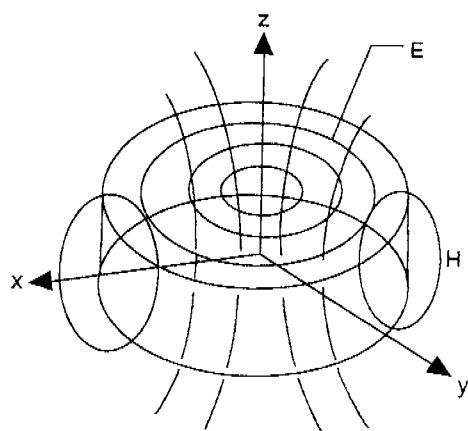
Figure 1. Field Distribution of $TE_{01\delta}$ Mode in a Dielectric Resonator
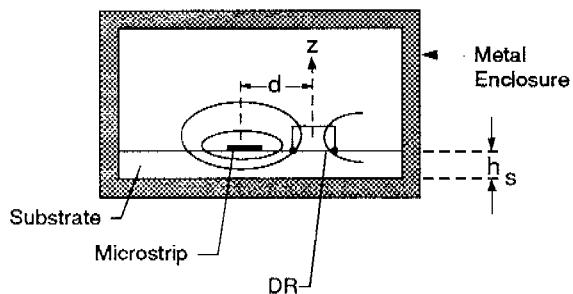
Figure 2. Dielectric Resonator Coupled to a Microstrip Line
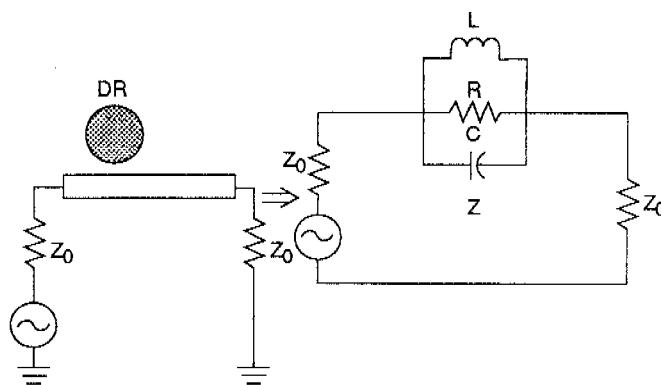
Figure 3 | Equivalent Circuit of D.R. Coupled
*& 4*| to a Microstrip Line
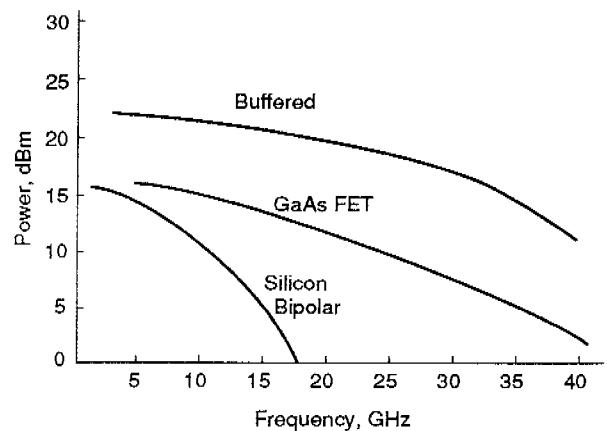
Figure 5. Power vs. Frequency for Transistor DROs

Table 1. Dielectric Resonator Materials

Composition	Dielectric constant	Q	Temperature coefficient of frequency (ppm)	Frequency range (GHz)	Manufacturer
$Ba_2Ti_9O_{20}$	40	10 000 at 4 GHz	+2	1–100	Bell Labs
$(Zr-Sn)TiO_4$	38	10 000 at 4 GHz	−4 to +10[a]	1–100	Trans Tech Thomson-CSF Murata
$Ba(0.33Zn0.67Ta)O_2$	30	10 000 at 10 GHz	0 ·10[a]	4–100	Murata
$Ba(0.33Mg0.67Ta)O_2$	25	25 000 at 10 GHz	4	4–100	Sumimoto
$BaO–PbO–Nd_2O_3–TiO_2$	88	5000 at 1 GHz	0 · 6[a]	<4	Murata-Trans Tech
Al_2O_3	11	50 000 at 10 GHz	0–6[a]	>18	NTK-Trans Tech

[a] Adjustable with composition.
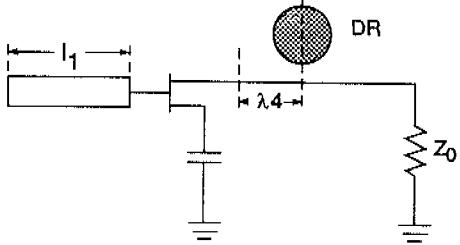
Figure 6. Stabilized GaAs FET DRO
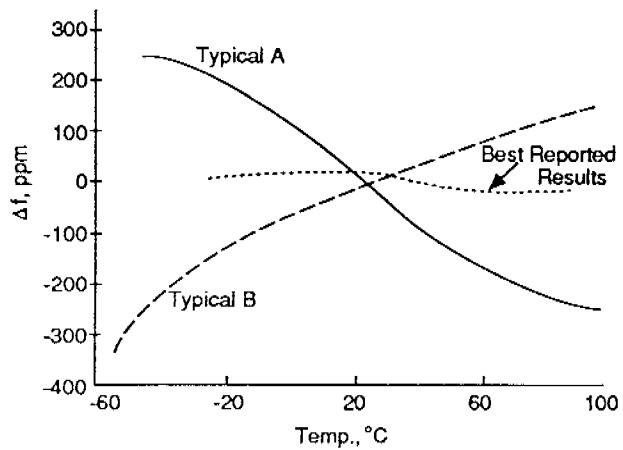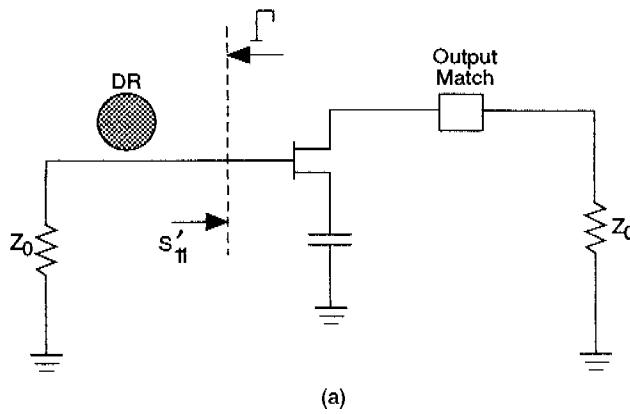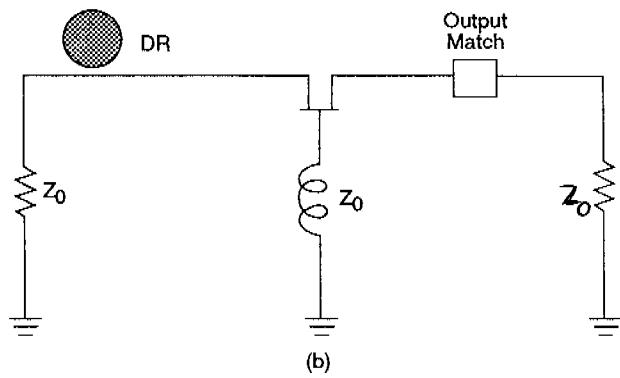
Figure 8. Frequency Drift vs. Temperature for
a Free Running 12 GHz GaAs FET DRO

Series Feedback
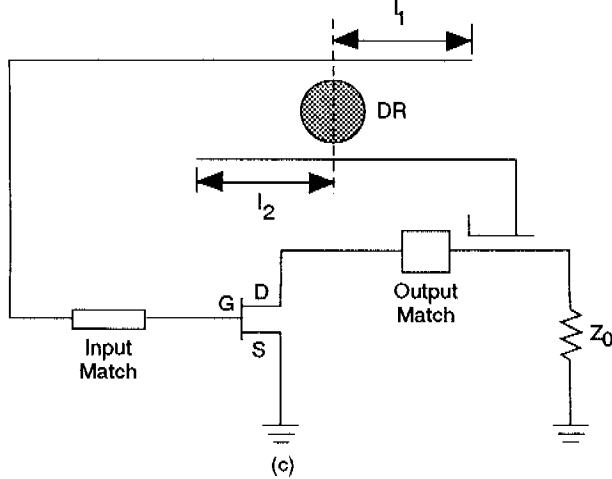
(a)
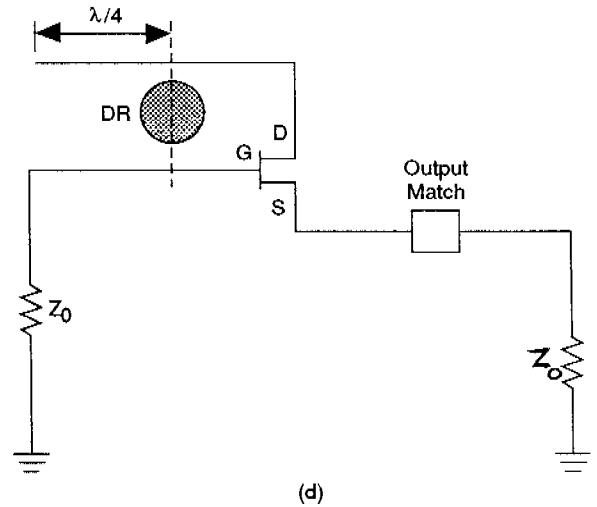
(b)

Parallel Feedback

(c)

(d)

Figure 7. Stable DRO Configurations
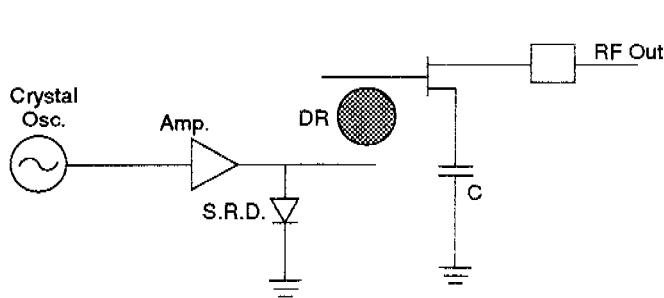
Figure 9a. Injection Locked DRO
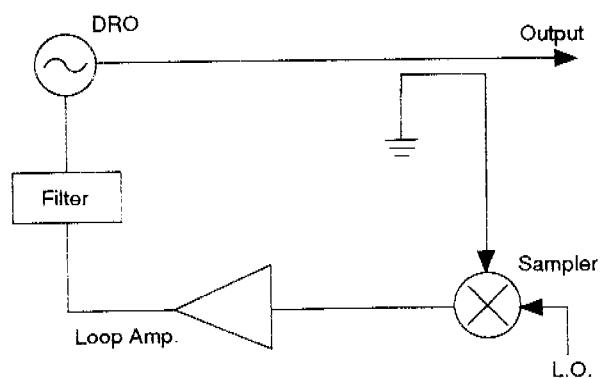

Figure 9b. Phase Locked DRO
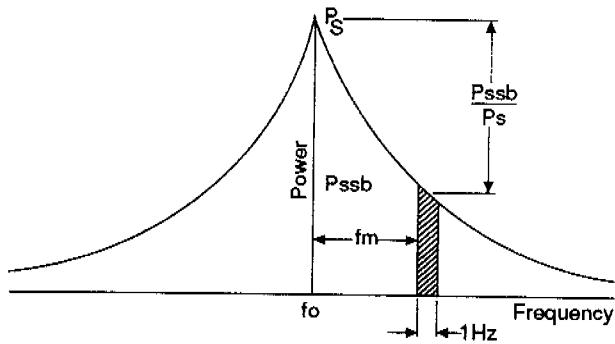

Figure 10. Oscillator Phase Noise
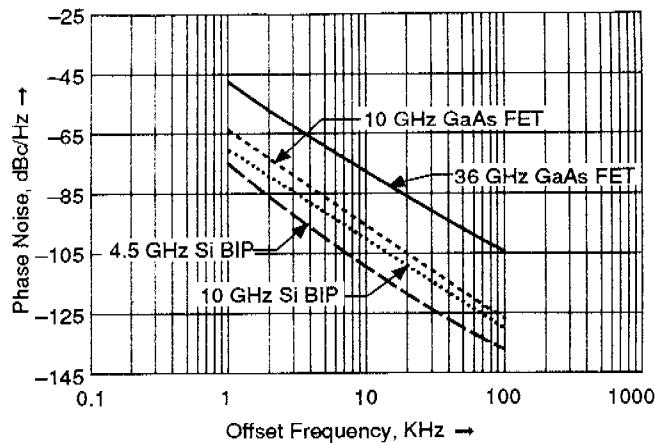
Figure 11. DRO Phase Noise:
GaAs FET & Bipolar
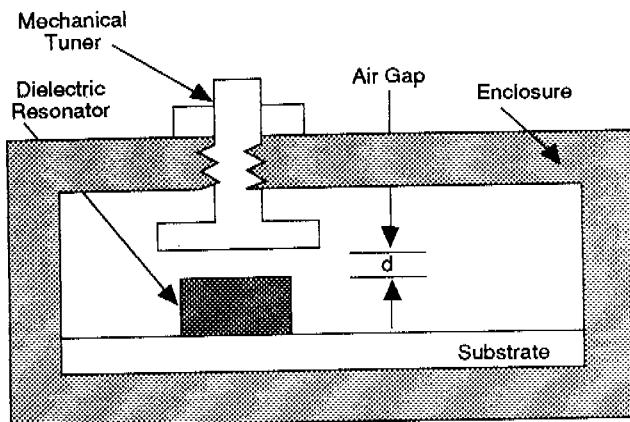

Figure 12. DRO Mechanical Tuning
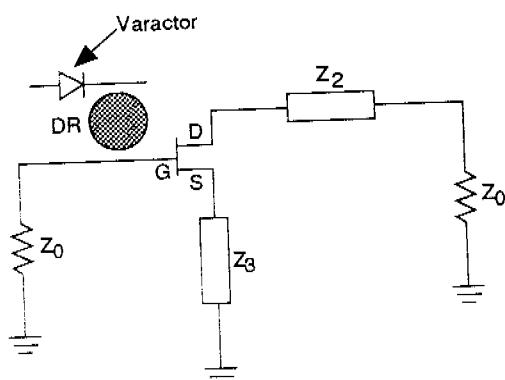
Figure 13. Varactor Tuned DRO

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QUESTIONS AND ANSWERS

JACQUES VANIER, NRC: Could you comment on the variation of Q with frequency?

MR. KHANNA: In the dielectric resonator, the $Q \times f$ product tends to be constant.