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REVIEW OF DIELECTRIC RESONATOR OSCILLATOR TECHNOLOGY

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Oscillators represent the basic microwave energy source for all microwave systems such as radars, communications, navigation or electronic warfare. A typical microwave oscillator consists of an active device (a diode or a transistor) and a passive frequency-determining resonant element. With the rapid advancement of technology, there has been an increasing need for better performance. The emphasis has been on low noise, small size, low cost, high efficiency, high temperature stability and reliability. Dielectric resonator oscillators offer the system designer a viable alternative in an effort to meet these challenges[1]. This paper will introduce the fundamentals of the transistor dielectric resonator oscillators, discuss various oscillator configurations, their performance, special circuits, limitations and future trends.

The quartz crystal oscillator is a highly stable source, however its operation is limited to less than 100 MHz. Stable microwave sources have commonly been realized using frequency multiplication of a the output of a quartz crystal oscillator. This method increases the FM noise power by N, where N is the multiplication factor, and has very low efficiency in addition to being very complex and expensive. Stable signals have also been generated in the past using metallic high Q cavities in passive cavity stabilization systems or in complex and bulky frequency discriminator systems.

The origin of solid state microwave oscillators using Gunn and Impatt diodes dates back to the late 1960s, before which microwave frequency generation depended upon klystron or magnetron tubes requiring massive power supplies. In less than two decades, solid state oscillators have evolved significantly. The extension of the bipolar transistor oscillator to microwave frequencies and the development of GaAs MESFET devices in the early 1970s has made available highly cost-effective, miniature, reliable and low-noise sources for use in the microwave and millimeter frequency ranges.

The basic active elements which can be used for microwave solid state oscillators are Gunn diodes, Impatt diodes and transistors. While Gunn oscillators have the advantage of low FM noise compared to Impatt oscillators, the latter has a higher efficiency and is a higher power device compared to the Gunn. Transistor oscillators on the other hand are low noise as well as high efficiency sources. Compared to Gunn oscillators, the transistor oscillators do not have the problems of threshold current, the necessity for heat sinking and the tendency to lock at spurious frequencies.

Gunn and Impatt diodes are negative-resistance devices, requiring only the application of a D.C. bias. Using them, the design of the oscillator is limited to the design of the output matching circuit in order to deliver the desired power output. Application of D.C. bias to the bipolar or the field-effect transistor, on the other hand, is not a sufficient condition for oscillation. Suitable series/parallel feedback is required to induce the negative resistance. The frequency range over which the negative resistance is present in the diodes is determined by the physical mechanisms in the device, while in the case of transistors this frequency range is also influenced by the chosen circuit topology. The only disadvantage to the transistor at present is a limit on the maximum oscillation frequency. While a Gunn or Impatt oscillator is capable of delivering 100 mW up to 100 GHz, the transistor oscillators are presently limited to approximately 10 mW at 40 GHz.

Transistor oscillators can be realized using either bipolar or GaAs FET devices. The maximum oscillation frequency for bipolar transistor oscillators is lower than that of GaAs FET oscillators. GaAs FET oscillators have been reported up to 100 GHz while oscillators using bipolar transistors have reached 20 GHz. However, the bipolar transistor offers lower phase noise: typically a bipolar oscillator has 6 to 10 dB less FM noise, (very close to the carrier) compared to a GaAs FET oscillator operating at the same frequency.

Dielectric resonators, due to their high Q, small size and excellent integrability in MIC circuits, can be directly used as the frequency determining element for realizing a MIC transistor oscillator. With the recent advent of temperature stable dielectric material, the transistor DRO is fast-becoming the most desirable choice in a vast number of applications.

The primary characteristics of the ceramic material to be used for dielectric resonators are:

- The Q factor, 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., possessing acceptable combinations of Q, τ_f and ε_r) severely limited dielectric resonator applications. Materials such as TiO_2 (rutile phase), 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 T_f value of 400 ppm/°C which makes it impractical for most applications.

The development of temperature-stable dielectric resonators dates back less than 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

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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 @ 10GHz, and a temperature coefficient which is variable through modification of the composition. Table I compares the important properties of different materials developed commercially.

the dual of a metallic cavity. The radiation losses of the dielectric resonators with the commonly used permittivities, however, are generally much greater than the energy losses in the metallic cavities, which makes proper shielding of the dielectric resonator a necessity. The 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 $1/\varepsilon_r$. If ε_r is high, the electric and magnetic fields are confined in region near the resonator, which results in small

TABLE I

DIELECTRIC RESONATOR MATERIALS

COMPOSITION	DIELECTRIC CONSTANT	Q	TEMPERATURE COEFFICIENT	FREQ. RANGE	MANUFACTURER
$Ba_2 Ti_9 O_{20}$	40	10,000 @ 4 GHz	+2	1 To 100 GHz	Bell Labs
$(Z_r - S_n) Ti O_4$	38	10,000 @ 4 GHz	-4 To 10 Adj	1 To 100 Ghz	Trans Tech Thomson, Murata
$Ba (Zn 1/3 Ta 2/3) O_2$	30	10,000 @ 10 GHz	0 to 10	4 To 100 Ghz	Murata
$Ba (Mg 1/3 Ta 2/3) O_2$	25	25,000 @ 10 GHz	4	4 To 100 GHz	Sumimoto
$Ba O - PbO - Nd_2 O_3 - Ti O_2$	88	5,000 @ 1 GHz	0 to 6	< 4 GHz	Murata/Trans Tech
$Al_2 O_3$	11	50,000 @ 10 GHz	0 to 6	> 18 GHz?	NTK/Trans Tech

Whether any of the dielectric compositions shown in Table I have overall superiority is not clear, since other factors, such as ease of ceramic processing and ability to hold tolerances on the dielectric properties must also be considered. The 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

The resonant frequency of a dielectric resonator is determined both by its dimensions and its surroundings. 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 certain resonant mode, such as $TE_{01\delta}$, can only be computed by rigorous numerical procedures. A number of theories on the subject, which can predict resonant frequency to an accuracy of $\pm$ 1% for the given configuration, appear in the literature. Unfortunately, these methods call for the use of high-powered computers. Kajfez[1] has presented an approximate solution of the involved equations both for the isolated case and for the more commonly-used MIC configuration. This method is typically accurate to $\pm$2%.

Currently-practical dielectric resonators cover a frequency range of 1 to 100 GHz. The lower frequency limit is imposed by the resulting large size of the resonator, while the upper frequency is limited by the reduced Q of small resonators, as well as by dimensions that become too small to effectively couple with the transmission line.

To a first approximation, a dielectric resonator is

radiation losses. The unloaded quality factor Q_u 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 parallelepiped shapes are also used. The commonly-used resonant mode in cylindrical resonators is denoted by $TE_{01\delta}$. In the $TE_{01\delta}$ 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.

To effectively use dielectric resonators in microwave circuits, it is necessary to have an accurate knowledge of the coupling between the resonator and different 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. The lateral distance between the resonator and the microstrip conductor primarily determines the amount of coupling between the resonator and the microstrip transmission line[2]. This has a direct effect upon output power, frequency stability, and harmonic content, as well as resonant frequency. 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 shows the equivalent circuit of the dielectric resonator coupled to the line.

DRO Circuits and Comparisons

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

A stabilized DRO is a device which utilizes a dielectric resonator in the output plane of a DRO circuitry to stabilize an otherwise free running oscillator (Fig. 4). 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, having 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 5 shows two common configurations of each type[1]. 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 double line coupling. In addition, the two coupling coefficients in the parallel case are not independent, increasing the difficulty of alignment. With the parallel 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 Parameters

Center Frequency and Power Output: As shown in Figure 6, Transistor DRO's are available spanning the frequency range of 3 GHz to 40 GHz with power outputs ranging to greater than +25 dBm. As noted earlier, the oscillator can be either Bipolar or GasFET, each with associated tradeoofsin performance, and the oscillator can be followed by one or mroe buffer amplifier stages as required to meet power output specifications. The center frequency is usually specified in MHz, to the MHz, with an associated stability (temperature dependent) or accuracy (temperature, pulling and pushing dependant) specification.

Frequency Stability is the measure of change in frequency over the specified operating temperature range. Typically, this has been expressed in parts per million per degree centigrade (ppm/^{o}C). As the actual frequency change versus temperature is not a linear function about a reference of +25^{o}C, other methods of specifying stability are preferred. One worthwhile method is to specify a percent change in frequency over a temperature range, another is to specify parts per million over the temperature range, referred to room temperature.

The principal cause of the DRO frequency drift with temperature 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, 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 D.R. with the microstrip line.
- The temperature coefficient (τ_{ph}) 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/^{o}C is generally required to offset the negative temperature coefficient of the device phase temperature coefficient. Figure 7 shows some of the typical frequency drift curves over temperature for free-running DROs.

With present technology it is now possible to repeatably produce DROs with frequency drift of less than $\pm$100 ppm over the -55 to +85 ^{o}C military temperature range, at frequencies up to 40 GHz. However, as shown in Fig. 7, results reported in R & D environment are significantly better ($\pm$ 10 ppm over temperature)[3].

Certain system applications require higher temperature stabilities than the ones achievable using free running DROs. A number of techniques used to improve the temperature stability are:

The Digitally Compensated DRO (DC-DRO), in which a temperature sensor is mounted in the oscillator to detect the temperature changes. Using EPROMs, pre-programmed with temperature characteristics of the DRO and a look-up table in the ROM, A/D & D/A convertors provide the necessary correction signal, which is applied to the varactor bias of an electronically-tuned DRO. Using this technique frequency stability of $\pm$ 15 ppm can be obtained over temperature[4].

The Analog-Compensated DRO (AC-DRO), which uses less-complex circuitry, consisting of an analog compensator circuit in conjunction with a temperature sensor, to achieve up to $\pm$20 ppm frequency stability. Each oscillator is characterized by generating a custom tuning voltage-vs.-temperature curve required to maintain a constant frequency. The compensation circuit is aligned to fit the tuning voltage vs. temperature curve of the specific oscillator. The correction voltage from the output circuit is applied to the tuning varactor of the ET-DRO, thus maintaining a constant frequency[4].

Ovenization is also 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 oC 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 better phase noise compared to the AC-DRO and DC-DRO because the DRO 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.

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 the previously-described approaches. A PL-DRO or IL-DRO approach also becomes necessary when multiple oscillators are required to be phase or frequency coherentor both. For locked systems, a highly stable crystal-controlled signal source operating at HF or VHF is used as a reference oscillator.

In injection 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 frequency of interest, and a free-running DRO is locked to the harmonic of the reference source through the injection locking circuit shown in Fig. 8. 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, Δf. This bandwidth is a function of the injection power, oscillator output power, and external Q.

Figure 9 shows a typical phase-locked DRO circuit. A DC-coupled sampler/phase detector is used to mix the nearest 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.

Injection locking is simpler and less expensive approach compared to phase locking, but the RF output is more likely to contain spurious signals at the harmonics of the reference oscillator frequency. Injection locking, in reality, is frequency locking as opposed to phase locking.

Single Sideband Phase Noise

Phase noise can be defined as short-term frequency stability, characterized by frequency variations in the output frequency which appear as FM energy around the carrier frequency. Phase noise is specified in dBc/Hz measured at specified offests from the carrier frequency: typical offsets are 10 and 100 kHz. In a DRO, phase noise is primarily dependent on four factors:

1. The low frequency noise inherent in the active device.
2. F_c, defined as 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 (or external Q) of the oscillator.

Optimization of phase noise performance calls for special design considerations, as well as the use of a high-Q dielectric resonator and proper device selection. 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. Figure 10 compares the phase noise performance of a number of different oscillators.

The following techniques can also be used to further reduce phase noise:

Low frequency feedback, using a parallel feedback circuit designed at low frequencies (up to 1 MHz) in order to reduce upconversion of the low frequency noise. Phase noise improvement of up to 20 dB has been reported using this technique[5].

Noise degeneration, using the same dielectric resonator both as the frequency determining element of the oscillator and the dispersive element of the frequency discriminator. The DC output of the discriminator is applied to the frequency control point of the DRO. This technique has been reported to achieve phase noise as low as -120 dBc @ 10 KHz at 10 GHz from the carrier[6].

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. 11). 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 magnetic walls is mostly magnetic, as is the case for the shielded TE_{01} dielectric resonator considered here, the resonant frequency will increase when the wall moves inward. Current designs allow for up to 5 percent tuning range without 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

The YIG-tuned oscillator (YTO) and varactor-tuned oscillator (VCO) are the commonly used electronically-tunable oscillators. YTO's are selected for high tuning linearity, large tuning bandwidth (>2 octaves) and low phase noise. Their disadvantages are low tuning speed, cumbersome mechanical size, considerable weight and significant tuning power consumption. VCO's, on the other hand, offer small size and weight, high tuning speed, low tuning power, but have relatively poor tuning linearity, inferior phase noise and generally lower bandwidth (< one octave). For some applications, however, such as FMCW radar sources, narrowband-modulated communication systems or PLL systems, bandwidth needs are on the order of 0.1 to 1% ; and these applications require sources with low phase noise, high tuning speed and low tuning power. Electronically tunable DRO's (ET-DRO) can now meet the requirements for many of these applications.

ET-DRO's are also commonly used for analog/digital temperature compensation of the oscillator. This application requires that the frequency tuning of the DRO should exceed 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 shown in Fig. 12) can provide up to 1% frequency adjustment. 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 sensitivity to changes in the supply voltage of the device used as an oscillator. By not using an internal voltage regulator, the oscillator can be designed to provide the necessary tuning range by varying the supply voltage, typically within .1% of center frequency. As output power is often a function of supply voltage, care must be exercised to maintain suitable output power variation characteristics.

Figure 13 shows the phase noise characteristics for an electronically-tuned DRO using a varactor. It may be noted that increase in electrical tuning results in increased phase noise. Figure 14 reflects the frequency tuning and modulation sensitivity characteristics with tuning voltage and also shows the output power fluctuation over the tuning range. The electrically tunable DROs are presently available up to 26.5 GHz.

In an optically-tunable dielectric resonator configuration (Fig. 15), photosensitive material, such as high-resistivity silicon, is placed directly on the dielectric resonator. Light from a laser or light-emitting diode (LED) is brought through an optical fiber to illuminate the photosensitive material, changing its conductivity and perturbing the electromagnetic fields in and around the resonator. This perturbation results in a shift in the center frequency of the DRO. Using this technique tuning bandwidths of >0.1% can be obtained.

Special DRO Circuits

Excellent integrability and high performance of the dielectric resonators has generated a large number of interesting DRO configurations for various system applications. Some of those special circuits will be described in this section:

A dual-resonator oscillator, shown in Fig. 16, presents a highly stable DRO circuit using identical resonators in a series feedback configuration in both the source and gate circuit of a FET[7]. This oscillator has three output ports, and its microstrip circuitry is wideband due to the simple, minimally-tuned 50-ohm lines on all three ports of the transistor. This oscillator is not particularly susceptible to spurious oscillations, and can provide high temperature stabilities by selecting appropriate dielectric mixes for the resonators.

A push-push DRO (Fig. 17) can use a common dielecric resonator for two transistors. In this approach the fundamental frequency of the oscillators is cancelled and the second harmonics are added at the output plane of the oscillator. This circuit helps to generate low-noise oscillations at frequencies much higher than otherwise possible. For example, using an 18 GHz resonator, a GaAs FET DRO was reported to have a phase noise of -100 dBc @ 100KHz[8]. Figure 18 shows a comparison of phase noise between an Avantek 18 GHz push-push bipolar DRO and fundamental GaAs FET oscillator.

A selectable multi-frequency oscillator development appears in Fig. 19 The diagram shows a fast-settling tri-frequency selectable DRO. A single GaAs FET is used in conjunction with a simple single-pole three-throw (SP3T) switch to select the dielectric resonator corresponding to the desired output RF frequency. Compared to the old approach of using several continuously-operating DROs and a high isolation, matched SP3T switch, this approach is free of the spurious signals at the unselected frequencies. This approach is also less expensive, uses less number of components and is more reliable compared to the old approach. The output frequency settles within ±10 ppm in less than 2 microseconds[9].

A silicon monolithic self-oscillating mixer using a Darlington pair is shown in Fig. 20. The dielectric resonator is used in parallel feedback between the input and output of the device. RF input is mixed with the LO in the first transistor Q1 and the second transistor offers gain at the IF frequency. Using this configuration an RF signal at 3.7 to 4.2 GHz was down converted to 1 to 1.5 GHz IF with 9 dB conversion gain [10].

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 effect of vibration on the performance of the DRO. Under vibration the variation of the distance between the resonator and the outer shielding is the major cause of the deterioration of the phase noise. A random vibration density of 0.2 G^2/Hz can increase the phase noise by more than 20 dB up to the highest frequency of vibration. Phase locking or injection 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.
- Extension of both lower and upper frequency coverage.
- Higher power at millimeter frequencies
- Reduction in cost and size
- Wider mechanical and electrical tuning bandwidths
- Improvement in DRO performance under vibration
- Optical tuning & injection locking
- Development of new materials for dielectric resonators for linear temperature coefficient and higher quality factor.
- Use of higher-order modes.

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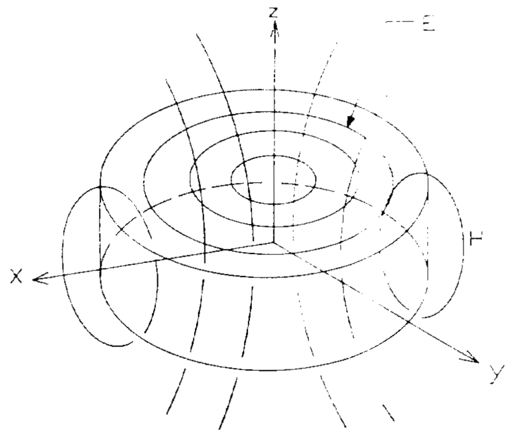


Fig. 1 Field distribution of TE_{01S} mode in a dielectric resonator.

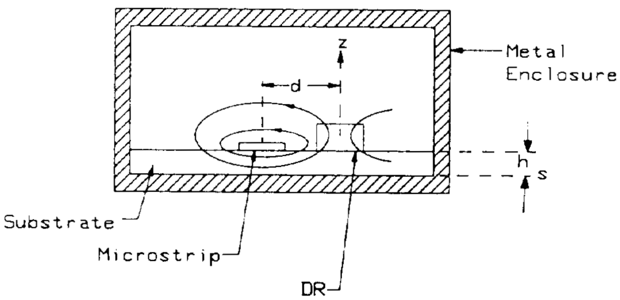


Fig. 2 Dielectric Resonator Coupled to a Microstrip line.

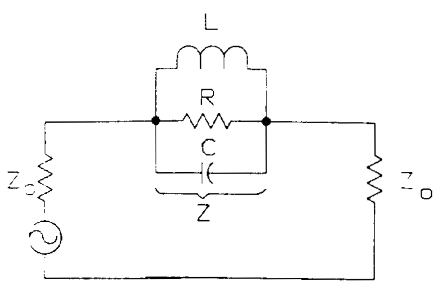


Fig. 3 Eqv. Ckt. of D.R. coupled to a microstrip line.

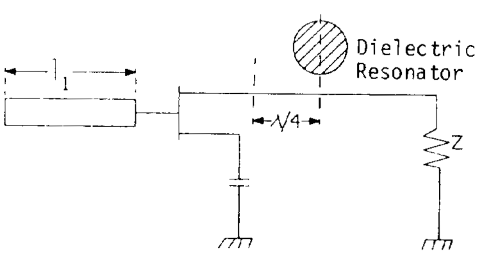


Fig. 4 Stabilized GaAs FET DRO.

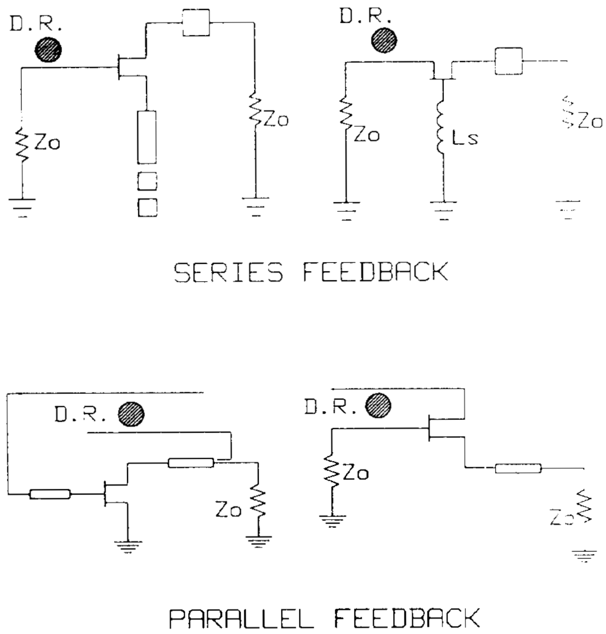


Fig. 5 Stable Transistor DRO Configurations.

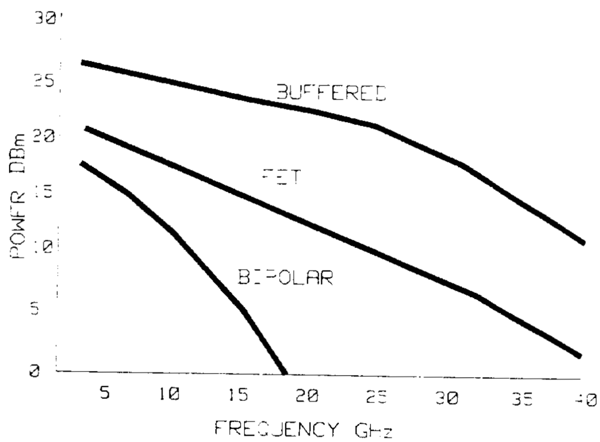


Fig. 6 Power vs. Frequency for Transistor DROs.

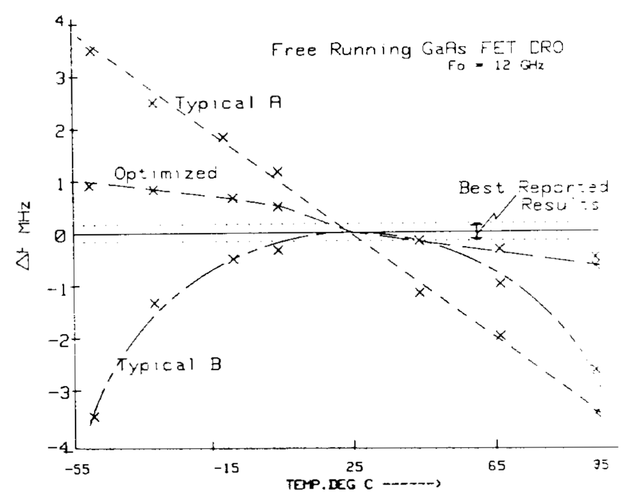


Fig. 7 Frequency Drift vs. Temperature for typical DROs.

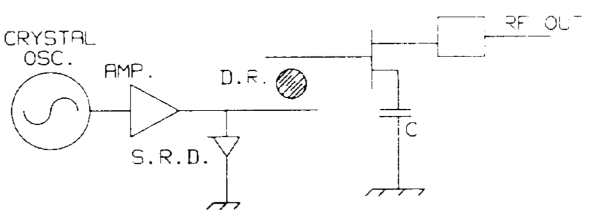


Fig. 8 Injection Locked DRO.

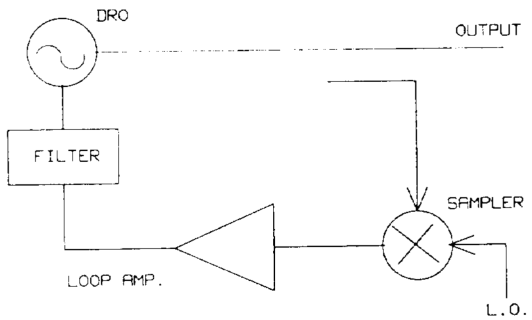


Fig. 9 Phase Locked DRO.

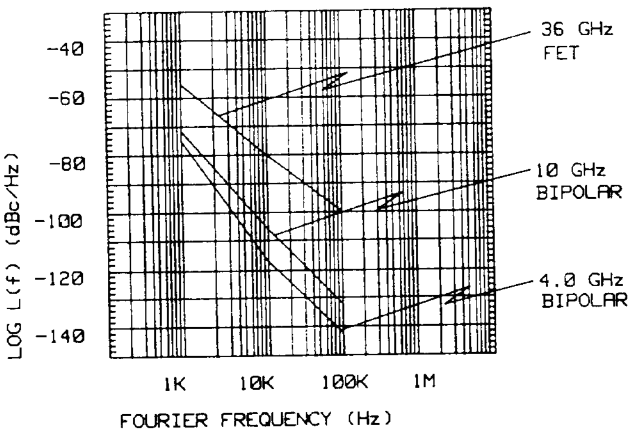


Fig. 10 Phase Noise of Transistor DROs.

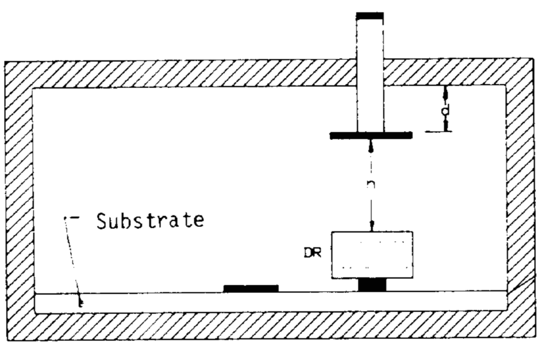


Fig. 11 Mechanically Tuned DRO Configuration.

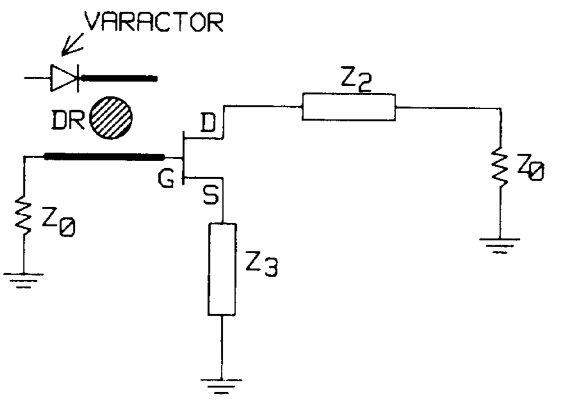


Fig. 12 Varactor Tuned DRO.

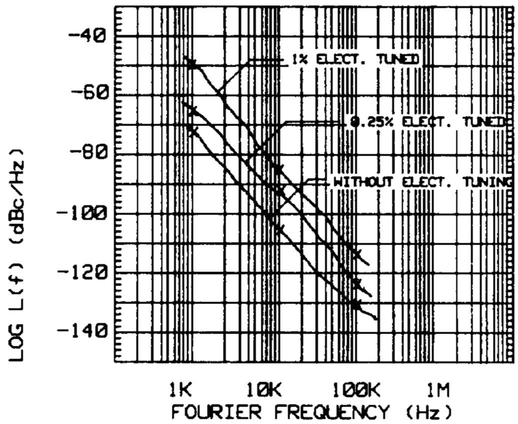


Fig. 13 Phase Noise Deterioration with Electronic Tuning.

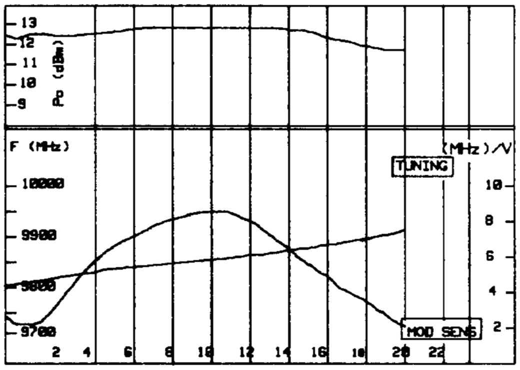


Fig. 14. Freq. tuning, modulation sensitivity & output power vs. tuning voltage for a VT-DRO.

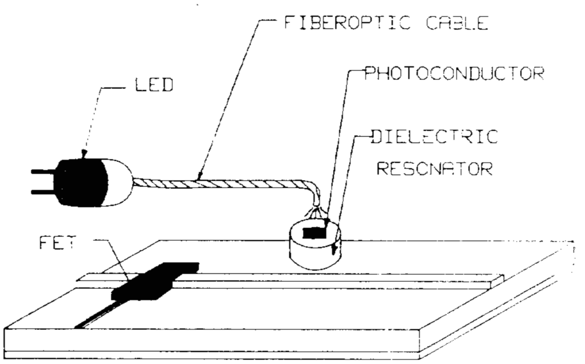


Fig. 15. Optically tuned Dielectric Resonator.

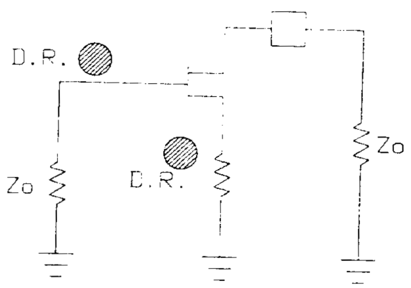


Fig. 16 Dual Dielectric Resonator Oscillator.

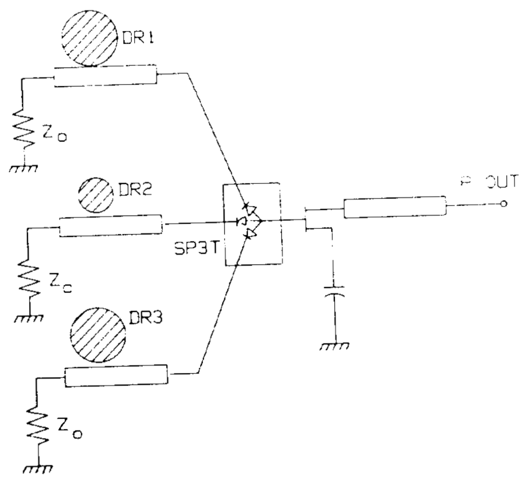


Fig. 19 Selectable Multi-frequency D.R.O.

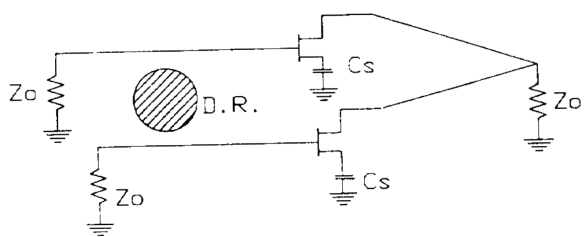
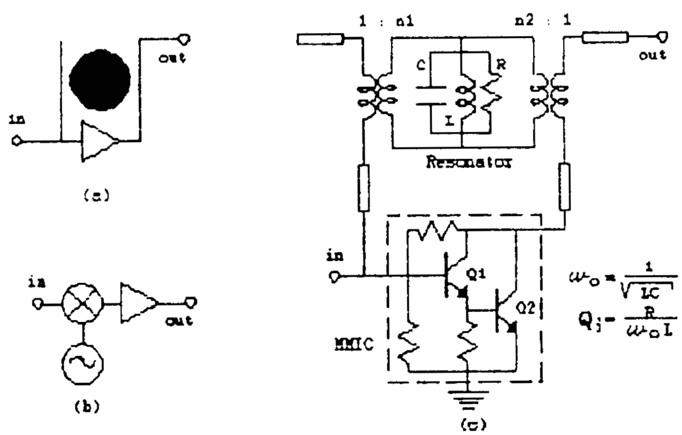


Fig. 17 Push Push Transistor DRO.



$\omega_o = \frac{1}{\sqrt{LC}}$

$Q_i = \frac{R}{\omega_o L}$

Fig. 20 D.R. Self Oscillating Mixer a)Schematic b)Functional Block Diagram and c)Equivalent Circuit.

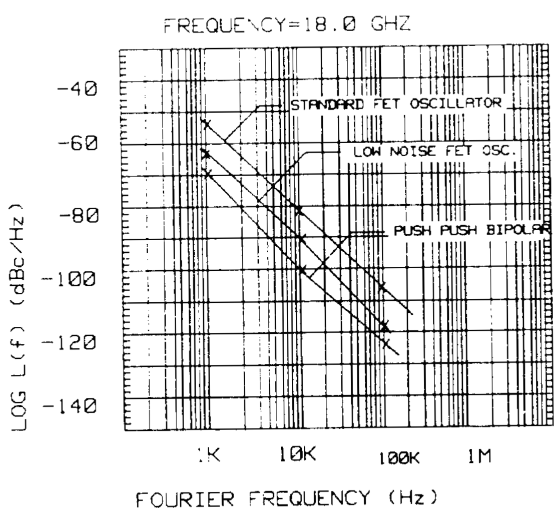


Fig. 18 Phase Noise Comparison between Push-Push Bipolar DRO & Fundamental GaAs FET DROs.