# An Ultra-Low Noise Microwave Synthesizer

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*Abstract*—A silicon bipolar transistor together with a barium titanate dielectric resonator were used to design a low noise microwave synthesizer. The oscillator was phase locked to a low-frequency (LF) reference with microwave frequency selection provided by a high-speed digital programmable divider within the phase-locked loop. The resulting FM noise $\Delta f_{rms}$ was 0.0003 Hz in a 1-Hz band greater than 1000 Hz from the 1-GHz carrier.

## I. Introduction

MICROWAVE generators with very-low near-carrier phase noise are required for space communication and tracking systems and for Doppler radar applications. These systems are sensitive to microwave generator phase noise in the range from several hertz to a few hundred kilohertz from the carrier. Most solid-state generators for such applications employ crystal oscillators in the 100-MHz range followed by frequency multiplier chains. To achieve the ultimate noise performance, two-cavity klystron oscillators are required along with their large power supplies and limited reliability. Both of these techniques represent a mature technology, the product of the last 15 years. Recent technology advancement in many areas suggests that the role of crystal oscillators in microwave sources requiring very-low near-carrier phase noise should be reexamined. We believe that in these applications a voltage controlled oscillator at 1 GHz offers an attractive alternative to a crystal oscillator and multiplier chain. To maintain long-term stability, the 1-GHz voltage-controlled oscillator (VCO) can be phase locked to a stable low-frequency reference with inexpensive digital circuitry. The 1-GHz signal can then be multiplied to the desired microwave frequency.

The microwave generator required for a communication system under development at Bell Telephone Laboratories (BTL) has stringent phase-noise requirements in the range of 200 Hz to 15 MHz from the carrier. These requirements could not be satisfied by any known existing commercial or Bell System product. In order to maintain the required long term stability (on the order of one part in $10^7$ per year), all microwave signals must be phase locked to a single extremely stable low-frequency reference signal. This paper reports the conclusion of our investigation of the design of a microwave generator for this system.

It is feasible to design a manufactureable 1-GHz all-solid-state frequency synthesizer with a frequency modulation (FM) noise ($\Delta f_{rms}$) less than 0.0007 Hz in a 1-Hz band at frequencies greater than 1000 Hz from the carrier. FM noise as low as 0.0003 Hz in a 1-Hz band 1000 Hz from the carrier has been achieved using this method. Realizing that oscillator noise is proportional to the frequency of oscillation, this result compares favorably to the lowest normalized noise ($\Delta f/f_0$), greater than 1000 Hz from the carrier, of any solid-state source reported in the literature. The work reported here was largely experimental in nature. It was however, guided by some relatively simple but extremely important concepts.

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## II. Oscillator Design Considerations

The design of the VCO used in the synthesizer was guided by the following well-known constraints. The FM noise of a free running oscillator is given by [1]

$$\Delta f_{rms} = \frac{f_0}{Q_E \cos\theta} \frac{kT_{eff}B}{\eta P_0} \tag{1}$$

where

- $f_0$ frequency of oscillation
- $\theta$ angle between a normal to the load line $Z(\omega)$ and the device line $Z(A)$
- $k$ Boltzmann's constant
- $B$ bandwidth
- $P$ the oscillator output power
- $Q_E$ the external $Q$ of the oscillator
- $\eta$ is a measure of the efficiency of the passive circuit$=1/(1+Q_E/Q_0)$
- $T_{eff}$ effective noise temperature of the oscillator.

Since $\Delta f_{rms}$ is proportional to $f_0$, a fundamental frequency transmission cavity oscillator and a transmission cavity oscillator with frequency $f_0/n$, followed by an ideal times $n$ multiplier will have identical FM noise, if the remaining quantities in (1) are unchanged. However, in practice, several factors favor the transmission cavity oscillator followed by a multiplier, when the lowest possible noise is required. As the frequency of oscillation increases, the output power available from an active device consequently decreases ($P_0 f_0^2=$constant), effective noise temperature of the active device increases and the intrinsic $Q$ of a cavity, $Q_0$, decreases. All of these factors provide a noise advantage for a transmission cavity oscillator followed by an ideal multiplier. A conventional crystal oscillator multiplier chain fails to take full advantage of these effects for the following reasons. Output power in a crystal oscillator is limited by large signal effects in the quartz crystal which cause $Q_0$ to decrease as $P_0$ increases beyond a given level. To minimize power dissipation in the crystal, a feedback oscillator configuration with a relatively low output power is commonly used.

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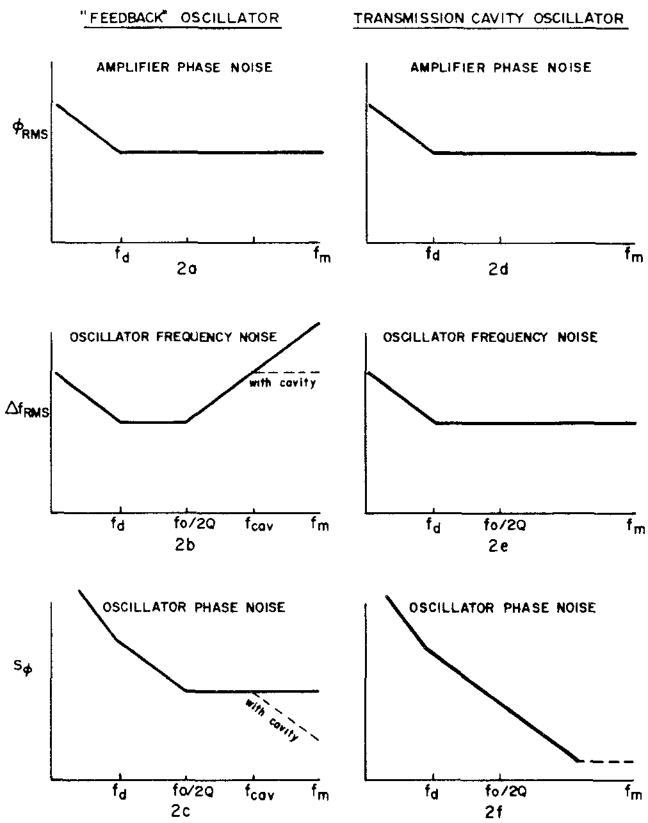


Fig. 1. Frequency and phase noise of an oscillator.

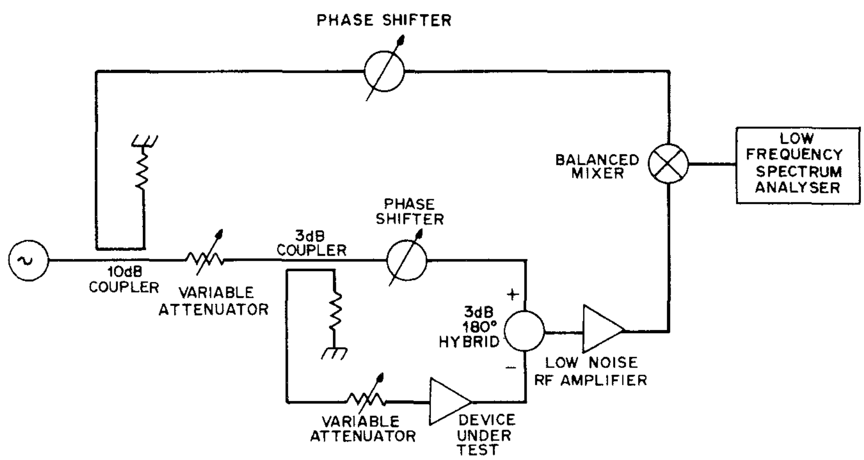


Fig. 2. Measurement setup for near-carrier noise of an amplifier.

As shown in Fig. 1, the feedback oscillator FM noise outside of the resonator passband increases with frequency. In this figure, $f_d$ is the $1/f$ noise corner of the active device and $f_{cav}$ is the 3-dB bandwidth of the secondary cavity. The level of the resulting phase noise floor is then increased by the subsequent multiplication process.

The effective noise temperature of the active device $T_{eff}$ is a function of both the frequency of oscillation (thermal and shot noise contributions) and the distance from the carrier ($1/f$ contribution). In addition both of these contributions depend upon the bias conditions of the active device. The sum of these contributions can be measured directly at microwave frequencies using the method described by Sann [2] and shown in Fig. 2.

In an effort to achieve very low FM noise over the range of 200 Hz to 15 MHz from the 6-GHz radio carrier we have chosen the following parameters for our design. The fundamental VCO is near 1 GHz followed by a broad band times 6 varactor multiplier. The VCO uses a high $Q$ temperature stable $Ba_2Ti_9O_{20}$ dielectric resonator, in a transmission cavity oscillator circuit with $\theta$ very nearly equal to 0.

The $1/f$ noise of the oscillator is determined by the active device characteristics. Therefore, selection of the proper device was very critical to the success of this generator. The $1/f$ noise is known to be a phenomenon produced by recombination within the device. This recombination causes fluctuations in the device current and voltage. This low frequency fluctuation can modulate the RF signal both in phase and amplitude. It is a linear modulation process for small RF signals but can be a nonlinear process for signal levels which are large enough to cause gain compression. This low frequency fluctuation effect also contributes to the rise in device noise figure at low frequencies. From the low frequency noise data, the Western Electric 82E n-p-n silicon bipolar transistor was chosen [3]. Individual devices were further selected for high dc gain at low emitter current. The dc $\beta$ of the device was required to be greater than 200 at an emitter current density of approximately 1 A/cm$^2$ with $V_{CE}$ equal to 10 V. This is a sensitive measure of base recombination and correlates well with low $1/f$ noise. The chip is then bonded into a common base microwave package with an integral stud for heat sinking the device. The optimum bias condition for this device was determined by measuring the near-carrier phase noise of the 1-GHz oscillator. This device is capable of delivering greater than the required 1 W of output power at 1 GHz. However, referring to (1), as the bias current level is increased to increase the maximum available output power, the resulting $T_{\text{eff}}$ near the carrier increases faster than the output power resulting in an increase in oscillator phase noise. The optimum noise performance was found at an emitter bias current density of about 500 A/cm$^2$. The low emitter current density chosen to optimize phase noise limited the RF output power and required the use of an RF amplifier. The additional phase noise from the amplifier will be negligible at frequencies near the carrier due to the large factor $f_0/Q_E$ in (1). Since the FM noise contribution of an amplifier is proportional to $f_m$ its effect will appear at high frequencies ($f_m > 2$ MHz for our generator).

## III. Synthesizer Circuit Description

A capacitive load consisting of a varactor in parallel with an adjustable capacitor is connected to the emitter of the Western Electric bipolar transistor. As shown in Fig. 3 this produces a negative resistance at the collector. The adjustable capacitor enables us to maximize the reflection coefficient at the collector $S'_{22}$ at the desired frequency of oscillation. This procedure in effect sets $\theta$ equal to 0 and in addition minimizes the output power variation with varactor voltage. Bias is supplied by a constant current source in the emitter and through the output 50-$\Omega$ transmission line to a voltage source in the collector.

As shown in Fig. 4, the 50-$\Omega$ transmission line from the collector is terminated at the far end with a lumped 50-$\Omega$ resistor. The output signal is extracted from an adjacent 50-$\Omega$ line through a dielectric resonator coupled to both transmission lines. The cylindrical dielectric sample, which is 2.1 in in diameter, is resonant in the $TE_{01\delta}$ mode. By adjusting the height of the resonator above the substrate and the spacing between lines, the proper external $Q$ and impedance transformation required for oscillation at the desired output power and frequency is provided.

The resonator is made of Barium Titanate ceramic ($Ba_2Ti_9O_{20}$) [4], [5]. This material has a relative dielectric

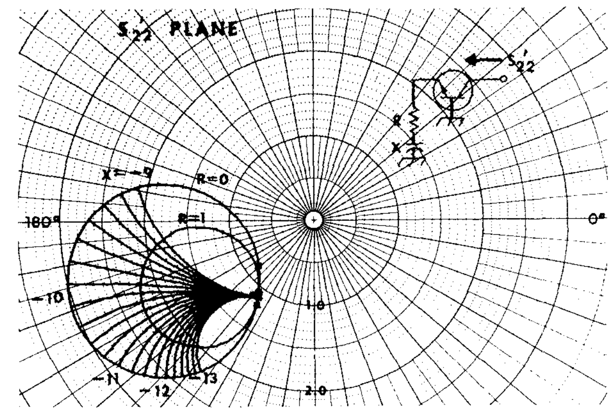


Fig. 3. Reflection coefficient at collector as function of emitter loading.

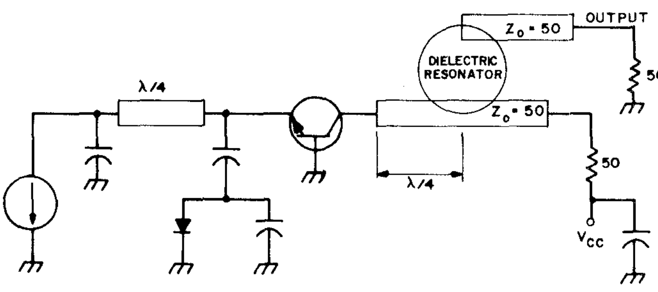


Fig. 4. Oscillator circuit.

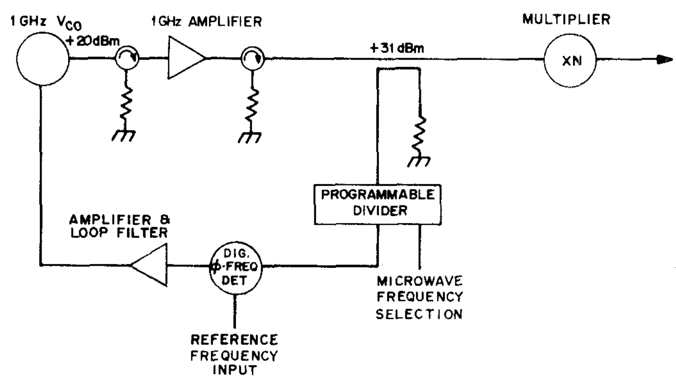


Fig. 5. Block diagram of the synthesizer.

constant of 39 and a temperature-frequency coefficient of 2 ppm/°C. At 1 GHz, the intrinsic $Q$ of the resonator is typically 23000 and decreases approximately linearly with increasing frequency. The external $Q$ of the oscillator circuit, determined from injection locking measurements, is about 7500. This choice of bias configuration and the resonator coupling arrangement eliminated all possibilities of spurious oscillation. The oscillator output power at 1 GHz is typically +19 dBm. A broad-band linear power amplifier with 12-dB gain, terminated by broad-band lumped element isolators, is printed on the microstrip circuit board along with the oscillator circuit. A sample of the 1-GHz amplifier output is obtained using a microstrip directional coupler. The sample is then applied to a high-speed ECL prescaler-programmable divider chain which divides the 1 GHz signal down to the frequency of the system reference signal. The output of the divider chain together with the system reference signal provide the input

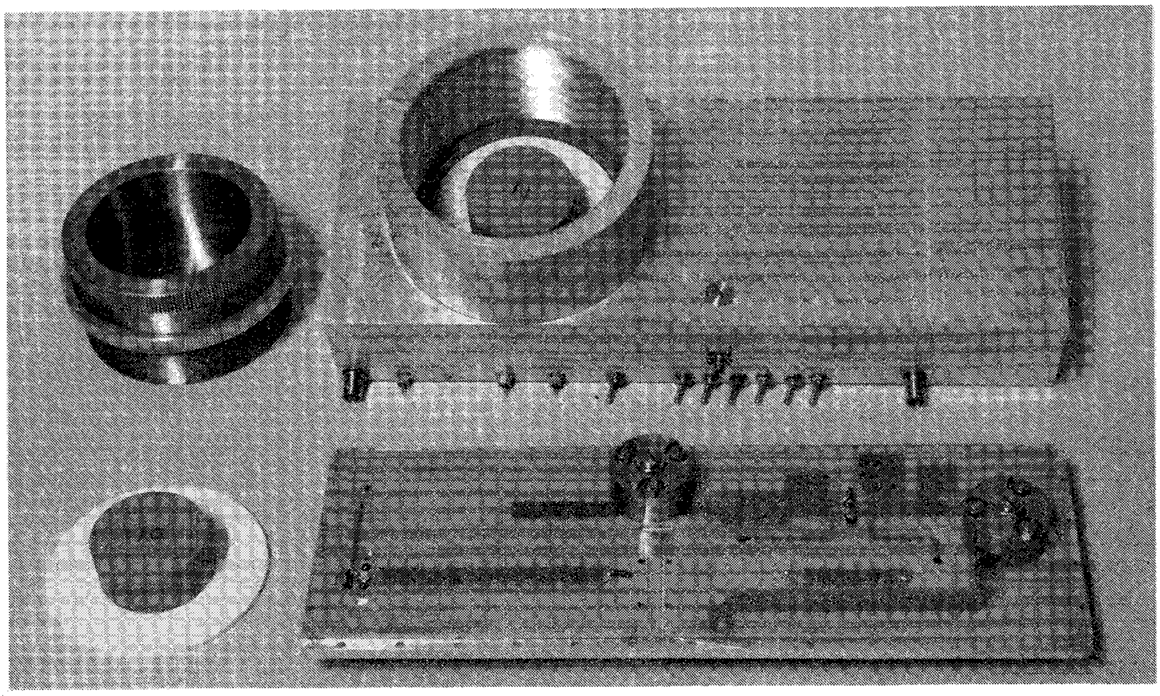

Fig. 6. Top view of synthesizer circuit.

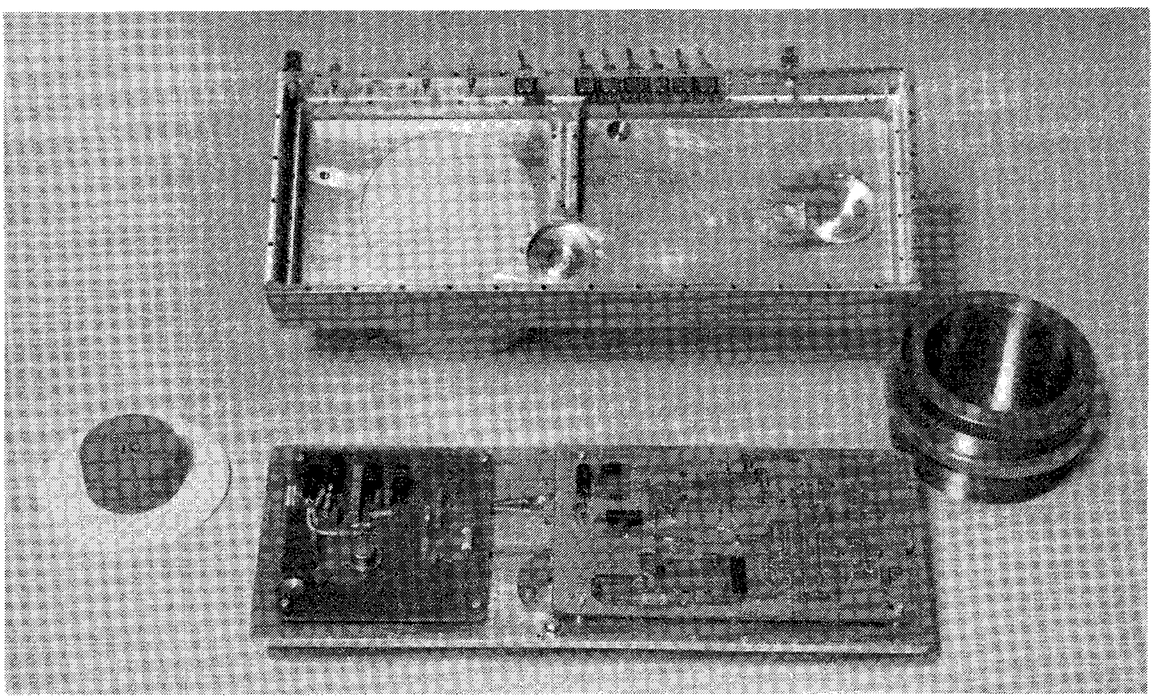

Fig. 7. Bottom view of synthesizer circuit.

to a digital phase-frequency detector. The detector output is filtered, amplified and applied to the varactor in the emitter of the oscillator transistor to close the phase locked loop. A block diagram of the synthesizer is shown in Fig. 5. A loop bandwidth of less than 1 Hz is used to prevent the noise of the reference signal from corrupting the oscillator phase at frequencies of interest. The desired microwave frequency is selected by setting the programmable divider to the required division ratio and adjusting the resonator frequency to provide the desired nominal varactor voltage. There are three circuit boards inside the aluminum housing. As shown in Figs. 6 and 7 all RF circuits are on one microstrip board. One printed wiring board is used for the phase locked loop circuits and another printed wiring board is used for the integral power regulator and bias network.

The dielectric resonator is supported by a ceramic disc inside a cylindrical cavity. The cavity dimensions provide tuning adjustments and prevent radiation loss. Tuning is accomplished by varying the fringing fields of the resonator with the metallic screw which forms the top wall of the cylinder. Although a range corresponding to the entire

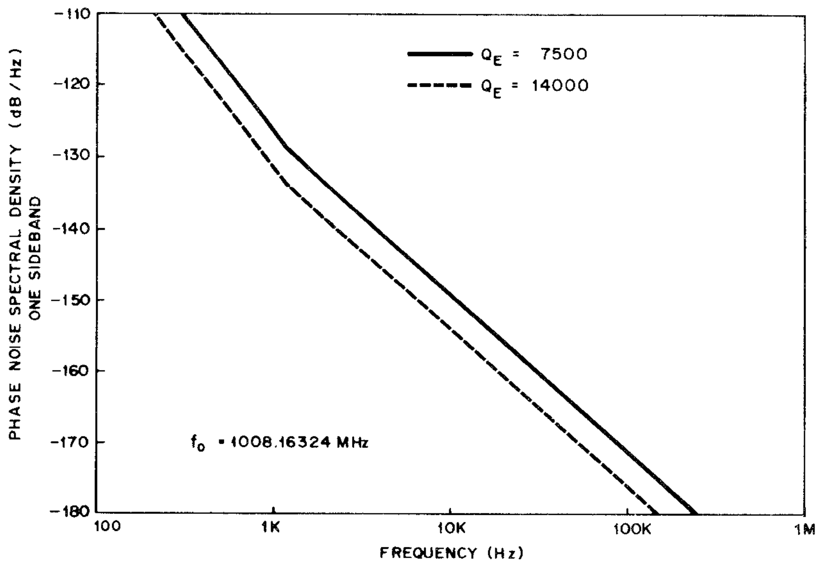

Fig. 8. Measured near-carrier phase noise.

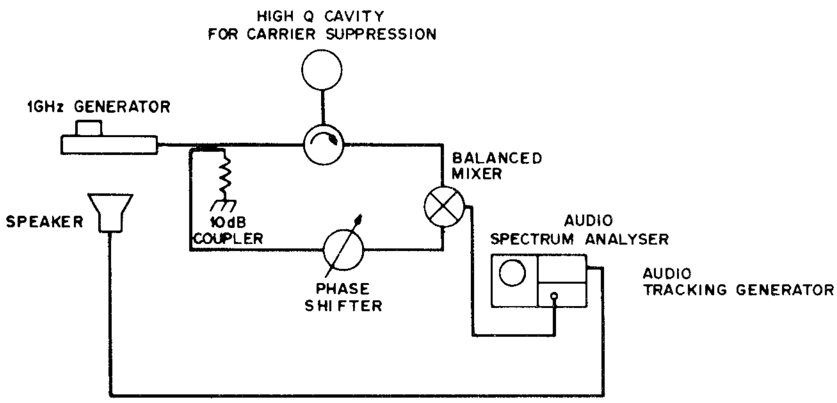

Fig. 9. Measurement setup for acoustic sensitivity.

6-GHz microwave radio band can be tuned with a single resonator, for optimum temperature performance a single resonator is recommended for each radio channel. The dielectric resonator has a temperature frequency coefficient of 2 ppm/°C, this temperature variation can be further compensated by the thermal expansion of the metallic housing. A properly compensated resonator exhibits a total frequency shift of $\pm 10$ ppm with an ambient temperature variation from 40 to 140°F. A typical varactor tuning range of $\pm 40$ ppm (with no more than 1-dB change in power) insures phase locking over this temperature range. In order to equalize the thermal time constants of this temperature compensation method the ceramic disk supporting the dielectric resonator is made of beryllium oxide.

## IV. Performance

The 1-GHz output power is typically +31 dBm. Sliding a 0-dB return loss load at the output through a phase of 360° produces a frequency change in an unlocked generator of less than 10 ppm. This is due to the high oscillator $Q_E$ and more than 80-dB isolation between the output load and the oscillator. With the loop locked all spurious tones are at least 120 dB below the carrier, and the phase noise is not affected by the system reference signal. The tone that can be identified is due to leakage of the 308-kHz reference signal and not from any spurious oscillation.

The spectral density of the oscillator phase noise was measured using the method described by Ondria [6] and Ashley [7] *et al.* and is plotted in Fig. 8. We have used a dielectric resonator with an external $Q$ of 23000 as the microwave cavity in this noise test set. During these tests the cavity was routinely required to absorb greater than 1 W of microwave power for long periods of time without any measureable degradation. By increasing the oscillator external $Q$ to 14000 laboratory measurements have repeatably shown a further 5-dB reduction in noise ($\Delta f_{\mathrm{rms}} <$ 0.0003 Hz above $f_m = 1000$ Hz). This is shown as the dashed curve in Fig. 8.

Because of its very low near-carrier phase noise, the effect of acoustic noise and mechanical vibration can be detected. The sensitivity of the phase noise to acoustic vibration is measured using the test set shown in Fig. 9 where a tracking generator tuned to the frequency of the spectrum analyzer is used to excite a loud speaker or shake table. Results of swept measurements using this technique are shown in Fig. 10 where dBC is a standard measure of acoustic pressure. For this scale 0 dB corre-

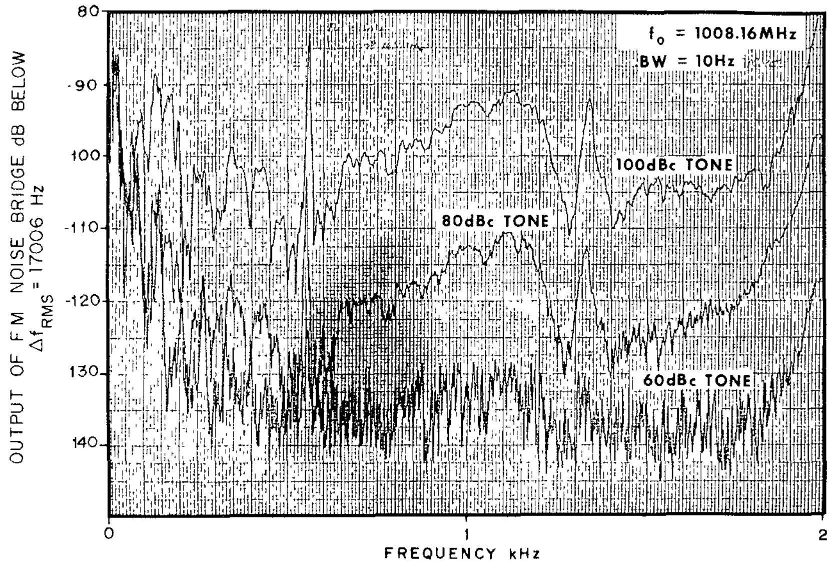


Fig. 10. Measured frequency "noise" as a function of swept acoustic tone level.

sponds to a pressure of $20 \times 10^{-6}$ N/m$^2$ and $C$ refers to $C$ weighting which is flat with frequency from 100 Hz to 3 kHz. With a single acoustic tone level below 60 dBC at the generator this effect is negligible, except at the resonant frequencies of the mechanical structure.

## V. Summary

We have designed and tested solid-state microwave synthesizers with very low near-carrier phase noise. The availability of low-loss temperature stable dielectric resonators make it possible to build a high-$Q$ resonant circuit. Because of the high power handling capability of these resonators, a transmission cavity can be used to improve the oscillator noise performance at frequencies beyond the resonator passband. The oscillator noise was further reduced by selecting devices with low $1/f$ noise and biasing these devices for optimum oscillator phase noise. The performance of the synthesizer is unexcelled, and because of the simple RF circuitry, inexpensive programmable digital phase-locked loop and lack of precision machined parts its manufacturing costs are potentially low.


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