Extremely Low Thermal Noise Floor, High Power Oscillators Using Surface Transverse Wave Devices

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***Abstract*—This paper presents state-of-the-art results on 1-GHz surface transverse wave (STW) oscillators running at extremely high loop power levels. The high-Q single-mode STW resonators used in these designs have an insertion loss of 3.6 dB, an unloaded Q of 8000, a residual PM noise of -142 dBc/Hz at a 1-Hz carrier offset, and operate at an incident power of up to $+31$ dBm in the loop. Other low-Q STW resonators and coupled resonator filters (CRF), with insertion losses in the 5–9 dB range, can conveniently handle power levels in excess of two Watts. These devices were incorporated into voltage controlled oscillators (VCO's) running from a 9.6-V dc source and provide an RF output power of $+23$ dBm at an RF/dc efficiency of 28%. Their tuning range was 750 kHz and the PM noise floor was -180 dBc/Hz. The oscillators, stabilized with the high-Q devices and using specially designed AB-class power amplifiers, delivered an output power of $+29$ dBm and exhibited a PM noise floor of -184 dBc/Hz and a 1-Hz phase noise level of -17 dBc/Hz. The 1-Hz phase noise level was improved to -33 dBc/Hz using a commercially available loop amplifier. In this case, the output power was $+22$ dBm. In all cases studied, the loop amplifier was found to be the factor limiting the close-to-carrier oscillator phase noise performance.**

I. Introduction

Surface acoustic wave (SAW) based oscillators are well known for their excellent phase noise performance in the frequency range of 0.1–1 GHz [1]–[3]. Resonator stabilized oscillators, operating in the 400–500 MHz range and featuring a phase modulation (PM) noise floor due to thermal noise of -184 to -185 dBc/Hz, were demonstrated recently [1], [3]. Along with a PM noise level at 1-Hz intercept of -48 to -55 dBc/Hz, these oscillators are considered to represent the current state-of-the-art phase noise performance in this frequency range. One limiting factor to further improvement of the oscillator noise floor is the power handling capability of the SAW device, which limits the maximum drive power level to about $+26$ dBm even if large area multitrack designs are used. An example was presented in [4] where a very high power SAW resonator failed after 200 weeks of operation at a power dissipation level of 130 mW (stress $= 200 \times 10^6$ N/m^2).

The fundamental power handling limit can be extended if the acoustic resonator uses the STW mode. As shown in [5], and experimentally verified in [6], devices with relatively small acoustic area can conveniently handle orders of magnitude higher drive power levels without degradation in performance. About a year ago, this unique feature was used to design a 1-GHz low noise voltage controlled oscillator (VCO) which was running at a loop power of $+34$ dBm and demonstrated a PM noise floor of -194 dBc/Hz [7]. This oscillator featured excellent tuning and wide-band frequency modulation capabilities. Its RF/dc efficiency was 19%.

In this paper, we present results from a one-year research effort on improved STW high power oscillators. Different STW resonant devices and 1-GHz fixed frequency and voltage controlled oscillators are characterized. The phase noise performance of the STW devices and oscillators, measured with different methods, is presented and discussed.

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II. STW Resonant Devices for Power Oscillator Applications

Compared to SAW resonators, STW devices offer a greater flexibility in the design of metal strip resonators and narrowband filters for oscillator applications. This increased design flexibility comes from the fact that metallization allows an additional degree of freedom in controlling the resonant Q while keeping low device insertion loss even in simple resonator configurations [8], [9]. This unique feature has been used extensively in the design of different kinds of single and multimode resonators and coupled resonator filters with a loaded Q, Q_L, ranging from 500–4400 and an insertion loss well below 10 dB at 1 GHz. All these devices can conveniently handle drive power levels in excess of two Watts and were found to operate without obvious performance degradation for several months in different fixed frequency and voltage controlled high power oscillators. The design details for such devices have been well documented in [5], [6], [10], and [11]. Here, we will characterize only some of the devices used in this study.

Fig. 1(a)–(d) presents data on a 1-GHz single mode high-Q resonator which was designed at the Institute of Solid State Physics in Sofia, Bulgaria, and fabricated using all quartz package (AQP) technology (AQP STW devices provided by G. K. Montress and T. E. Parker). This device has an insertion loss of 3.6 dB, a loaded Q of 2740, and an unloaded Q of 8000. It was intended for use in a fixed frequency power oscillator. However, if run at a loaded Q of about 3000, a tuning range of about 180 kHz over the 1-dB device bandwidth should be possible [see Fig. 1(b)].

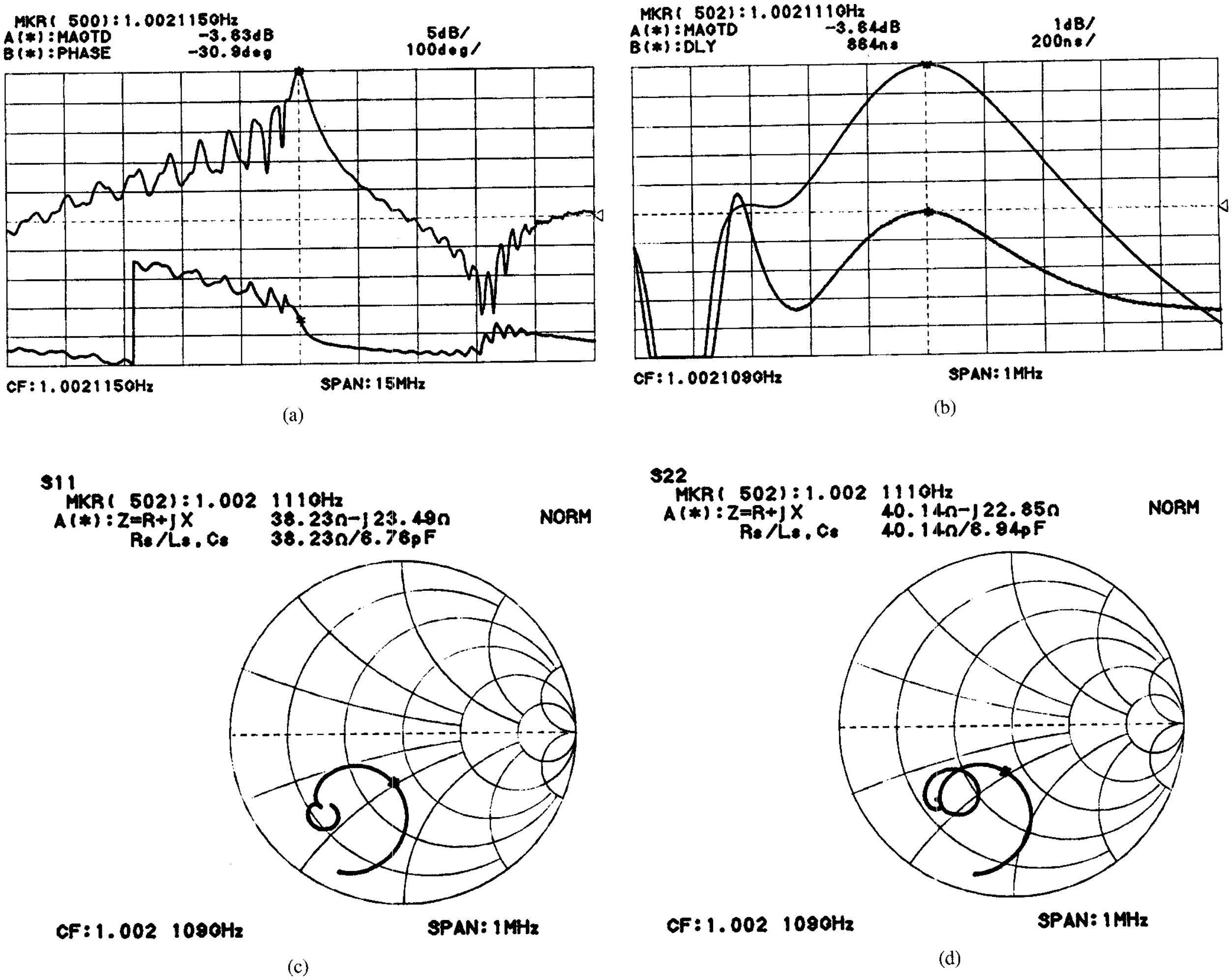

Fig. 1. Characteristics of a single mode high-Q AQP STW device. (a) Frequency and phase responses. (b) Detailed frequency and group delay responses. (c) Input reflection coefficient S11. (d) Output reflection coefficient S22.

A much wider tuning range can be achieved with low-Q resonators. Design techniques for obtaining lower device Q are described in [6], [10], and [11]. The device, characterized in Fig. 2, has an insertion loss of 5.2 dB, a loaded Q of 1500, and allows a tuning range of about 700 kHz when the VCO is tuned over the 3-dB device bandwidth.

If even wider tuning ranges are necessary, the two-pole coupled resonator filter, characterized in Fig. 3, can be used. This device has an insertion loss of 8.5 dB and a 1-dB bandwidth of 1.5 MHz. As is evident from its phase response, a variable phase shift of 0–180° would be necessary for tuning over the entire 1-dB bandwidth. In this case, three cascaded C-L-C varactor tuned phase shifters could be used [4].

All devices were fabricated in a standard single-step photolithographic process with careful control over the metallization. The resolution required was about 1.3 μm, which can readily be realized with almost all kinds of photolithographic equipment available to date.

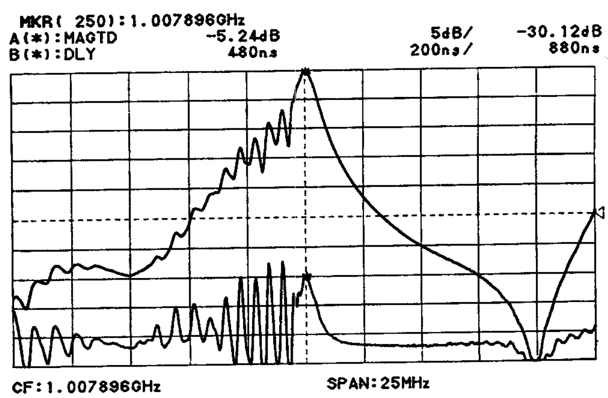

Fig. 2. Frequency and group delay responses of a low-Q STW resonator with a loaded-Q of 1500 and an insertion loss of 5.2 dB.

III. STW High Loop Power Oscillator Designs

As demonstrated with the STW oscillator described in [7], a substantial improvement in the thermal noise floor can be achieved if the loop amplifier is capable of generating output power levels in excess of two Watts. At GHz frequencies, these levels are very difficult to achieve with A-class amplifiers using bipolar transistors which are known for their low $1/f$ noise. Although some expensive power transistors can generate

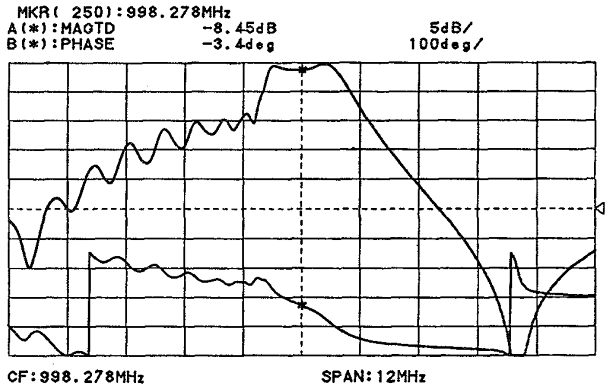

Fig. 3. Frequency and phase responses of a two-pole coupled resonator filter with a 1-dB bandwidth of 1.5 MHz.

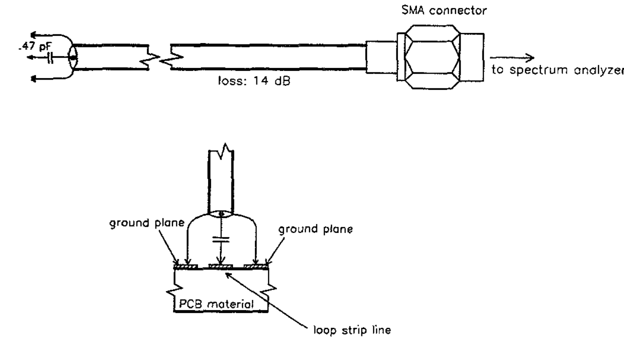

Fig. 4. A capacitive probe for evaluation of the oscillator loop power under closed-loop conditions.

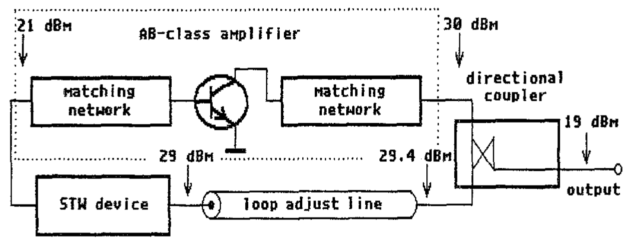

Fig. 5. Simple STW high loop power oscillator with directional coupling to the load. Block and level diagram.

a high output power in the A-class of operation and a 50-Ω environment, their efficiency rarely exceeds 10%. An elegant and inexpensive solution to the efficiency problem can be obtained if AB-, B-, or C-class amplifiers are used. With one of the STW oscillators, using an AB-class loop amplifier [7], we were able to achieve an RF/dc efficiency of 36% with an oscillator output power of +28 dBm at 1 GHz. One major drawback of AB-class amplifiers is that they require careful reactive matching at their input and output in order to be able to work efficiently in a 50-Ω environment. This is because the input and output impedances of the power transistors in the AB-class of operation are up to an order of magnitude lower than 50 Ω. If such an amplifier is to be used in an STW power oscillator, the matching circuits under closed-loop conditions have to be changed in order to achieve a low reflection coefficient at the input and output of the STW device since its impedances also differ from 50 Ω [see Fig. 1(c), (d)]. Under these circumstances, measuring the oscillator's loop power is a serious problem. Just breaking the loop open and loading it on both sides with the 50-Ω impedances of the measurement system will not work because this will seriously deteriorate matching. Therefore, the loop power has to be measured under closed-loop conditions. We have solved this problem by means of the capacitive probe shown in Fig. 4. It consists of a piece of coaxial cable ending with a small 0.47-pF capacitor soldered in series with the center conductor. The ground skirt is split into two parts, symmetrically bent on both sides of the cable in such a manner that the probe can conveniently touch any point of the loop's strip lines and ground planes, surrounding them, as shown in Fig. 4. The reading is obtained from a spectrum analyzer or power meter connected to the other end of the cable. The probe is calibrated by touching the load to the oscillator's output where the power can be precisely measured with a power meter. This reading will give the attenuation of the probe (14 dB, in our case). The loop power at any other point of the loop is obtained by adding the probe loss to the reading. We found that the probe did not deteriorate the matching conditions at the points of measurement. Only a slight frequency shift of up to 20 ppm was observed. This shift was well within the tuning range over which the oscillator's output amplitude was observed to be constant.

We investigated three types of high loop power STW oscillators using highly efficient AB-class loop amplifiers. Since the operating principle of such oscillators was described in detail in [7], we will present only the block circuits and loop level diagrams measured with the capacitive probe.

The simplest circuit, which requires a minimum number of passive components and only one power transistor, is shown in Fig. 5. It was designed to run with minimum loss around the loop. This was achieved using an 11-dB directional coupler instead of a 3-dB power splitter for load coupling. Thus, the loss (typically 3.5–4 dB) of a 3-dB power divider was reduced to 0.6 dB. With a loop power of +30 dBm and an output power of +19 dBm, the oscillator was found to provide stable fixed frequency operation and was insensitive to load changes. Unfortunately, we were unable to achieve usable frequency tuning with this design. The use of a varactor tuned phase shifter in the loop made the oscillator unstable due to a deterioration of the matching conditions. Another drawback of this design was that the gain compression had to be kept very low (1–2 dB) because of the poor limiting action of the AB-class amplifier. Increasing the loop power results in an increase of the collector current which can thermally overload the transistor.

The tuning and limiting problem was readily solved with the circuit in Fig. 6. Here, an A-class amplifier is incorporated between the variable phase shifter (VPS) and the AB-class power amplifier. Its function is twofold. First, it provides isolation between the VPS and the power stage, and second, it performs the limiting function, thereby providing a safe input power level to the power transistor. Thus excellent tuning

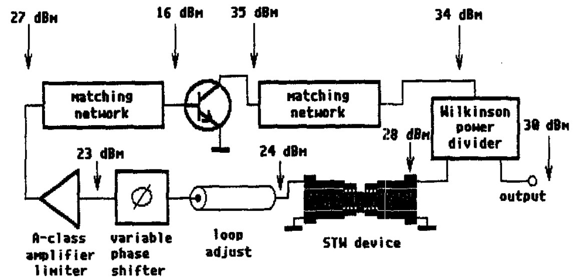

Fig. 6. Block and level diagram of a 1-W VCO stabilized with the high Q device characterized in Fig. 1.

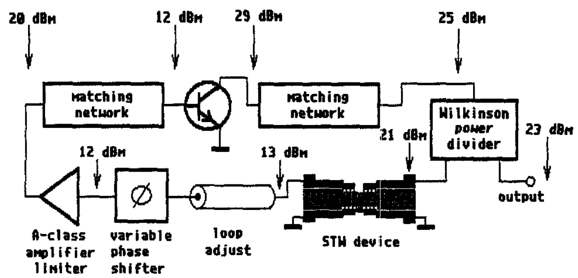

Fig. 7. A 9.6-V VCO for portable applications with RF to dc efficiency of 28% and an output power of 200 mW.

over the 3-dB device bandwidth and stable amplitude over the entire tuning range are assured. However, one has to be careful when adjusting the gain compression. Too much gain compression will result in a deterioration of the overall phase noise performance, as is evident from the phase noise plot in Fig. 19 which will be discussed later in this paper. A trade-off between gain compression, loop loss, device Q, and tuning range can be achieved by adding series capacitors to the STW device. This increases its insertion loss and loaded Q. In this case, a readjustment of the matching circuits is necessary. Another 1–2 dB variation of the gain compression is also possible by unbalancing the Wilkinson power divider. This changes the output power accordingly.

The circuit in Fig. 6 was stabilized with the high-Q device from Fig. 1. The loop power was measured to be +35 dBm and the incident power on the STW device could be altered between +28 and +31 dBm by unbalancing the power divider. According to Leeson's model, this oscillator should have a PM noise floor of −195 dBc/Hz due to thermal noise [1]. The amplitude modulation (AM) noise floor due to thermal noise should be similar to the PM noise floor. However, AM noise levels were not measured on these oscillators.

Fig. 7 shows a highly efficient high power VCO which uses the same concept. Since a wide tuning range of 700 kHz was necessary, the oscillator was stabilized with the low-Q resonator from Fig. 2. It was designed to run from a 9.6-V dc rechargeable NiCd battery for portable applications. The output power is +23 dBm and the RF/dc efficiency is 28%. When run at a supply voltage of 16 V, the output power increased to +28 dBm and the RF/dc efficiency decreased by only 3%.

Fig. 8, curves (A)–(C) shows the tuning characteristics of the high power VCO's from Figs. 6 and 7. They all were obtained with a single C-L-C type VPS which was designed to deliver about 60° of variable phase shift [4]. Cascading two identical phase shifters and using the CRF from Fig. 3 would increase the tuning range to 1.5 MHz with a tuning voltage of 0–9.6 V.

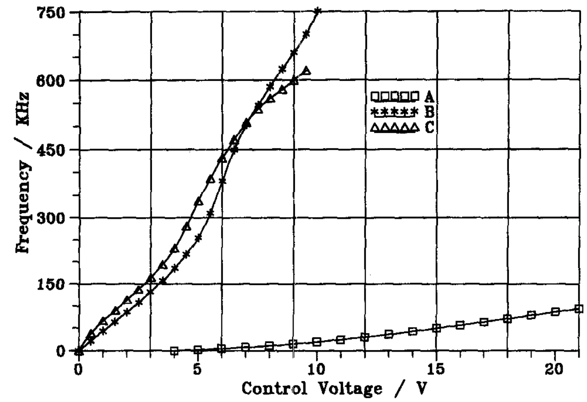

Fig. 8. Tuning characteristics of the power VCO's: (A) A 1-W VCO using the high-Q device from Fig. 1. (B) A 9.6-V wide tuning range VCO with 28% efficiency. (C) A wide tuning range VCO at 16-V supply voltage and +28-dBm output power.

IV. Phase Noise Measurements

The major goal of this study was to evaluate the phase noise performance of STW power oscillators at 1 GHz since this is important to a variety of applications in this frequency range. Since it is difficult to obtain the overall oscillator phase noise data for Fourier frequencies ranging from 1 Hz to 10 MHz away from the carrier with one single measurement, especially if two identical oscillators are not available, we set up different measurement systems and performed several measurements on the high power oscillators to make sure that the various systems delivered comparable results.

First, we tried to measure the oscillator PM noise floor. The simplest and most forgiving system for phase noise floor evaluation is the single-channel frequency discriminator with coaxial cable of delay τ_g, shown in Fig. 9. This system is not sufficiently sensitive for close-to-carrier measurements on stable oscillators but provides very good results for Fourier frequencies as high as 35% of the frequency at which the first null of its transfer function occurs ($f < 0.35/\tau_g$) [12]. Moreover, it requires only one oscillator and adapts to small changes of the oscillator's frequency during the measurement.

If the outputs of two identical channels of this system are cross-correlated in a two-channel FFT analyzer (Fig. 10), then the uncorrelated system noise averages towards zero as $1/\sqrt{n}$, where n is the number of averages. The system sensitivity can usually be improved by 20–25 dB [12]–[14] by using this approach. This cross-correlation concept can also be applied to a measurement system using two identical oscillators (Fig. 11). The system phase noise floor in this case can be better than −195 dBc/Hz, even at very high Fourier frequencies [12]–[14].

Residual noise measurements on the high-Q STW device from Fig. 1 were performed with the setup in Fig. 12, adapted from [4]. A low PM noise frequency synthesizer (HP 8662

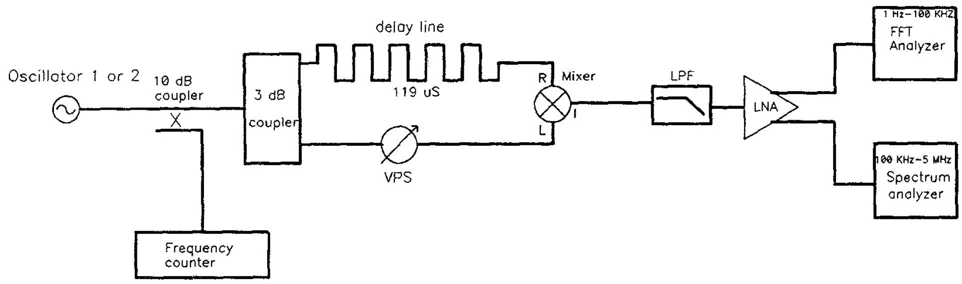

Fig. 9. Measurement setup using a single-channel delay line frequency discriminator.

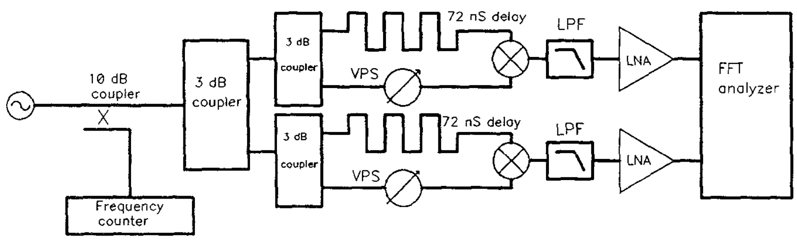

Fig. 10. Cross-correlation frequency discriminator measurement setup.

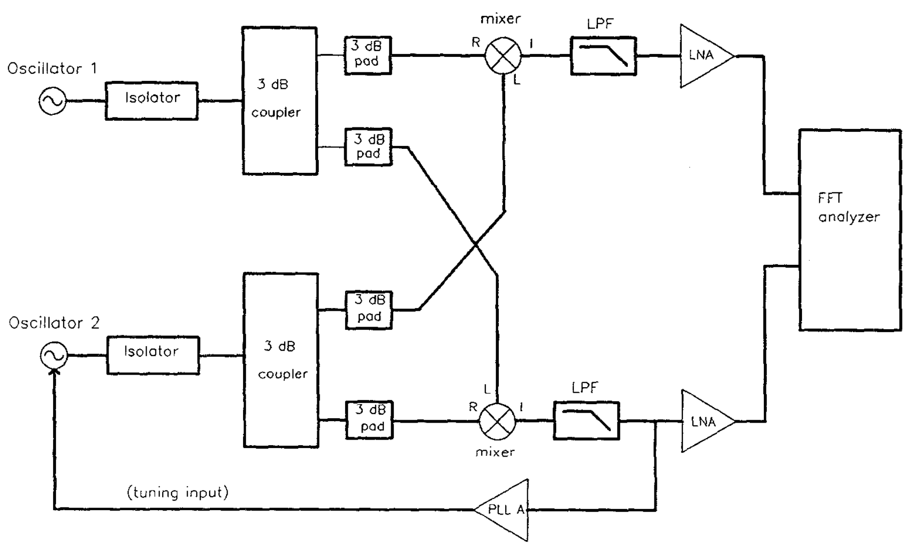

Fig. 11. Two-oscillator cross-correlation measurement.

A) was used to make phase noise measurements for frequency offsets from 2 Hz to 1 kHz from the carrier. To provide high enough mixer drive level and sufficient suppression of the synthesizer AM noise, we amplified the source signal with two cascaded low PM noise amplifiers (UTO 1023, Avantek), the second of which was driven into 3 dB of gain compression.

Close-to-carrier phase noise measurements on very low-noise STW oscillators using the low PM noise amplifier (UTO 1023, Avantek) were performed with the system setup in Fig. 13. This setup requires that the reference source (in this case, a low-noise frequency synthesizer) be substantially quieter than the STW oscillator. In our case, this approach does not hold for offsets greater than 1 kHz.

V. Evaluation of the Phase Noise Data

Fig. 14 shows the phase noise plot of two nearly identical power oscillators using the design in Fig. 6. The loaded Q of the STW devices was adjusted to a value of about 4000, which is half of the unloaded Q. The output power in this case was +29 dBm. PM noise floors of -184 and -182 dBc/Hz were measured with the single-channel frequency

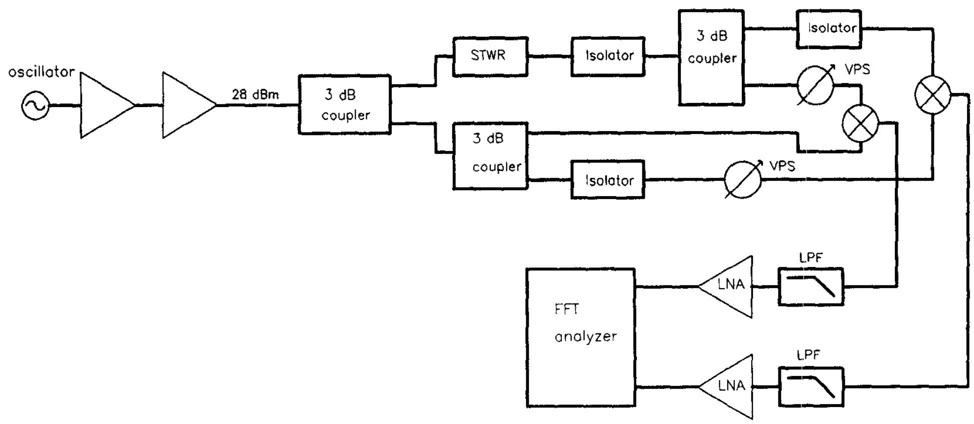

Fig. 12. Setup for residual phase noise measurements on high-Q STW devices.

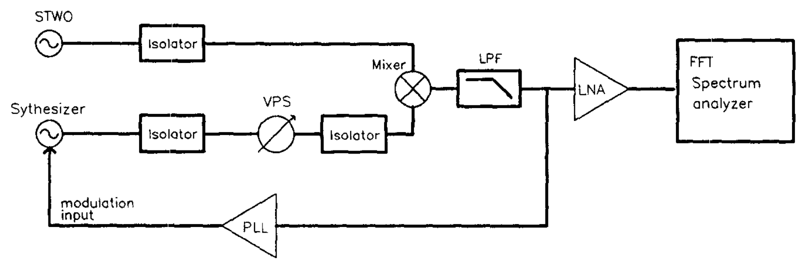

Fig. 13. System for measuring the close-to-carrier phase noise of fixed frequency STW oscillators using the low PM noise power amplifier (UTO 1023, Avantek).

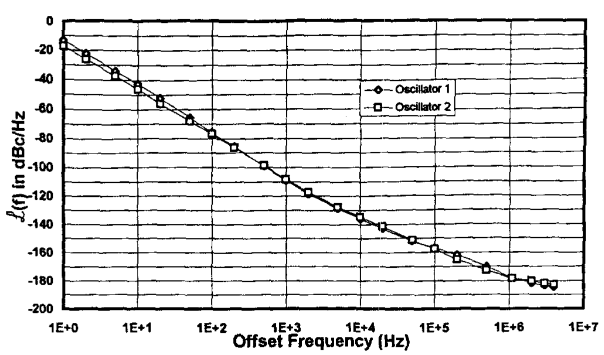

Fig. 14. Phase noise plots of two high-Q STW power oscillators measured with the single-channel frequency discriminator method. Both oscillators have $Q_L = 4000$, $P_{\text{out}} = +29$ dBm.

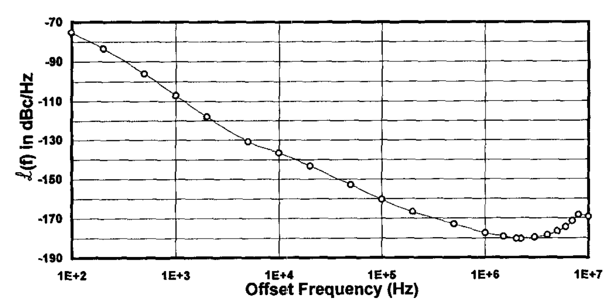

Fig. 15. Phase noise data of the high-Q oscillators obtained from the two-oscillator cross-correlation measurement (Fig. 11) assuming equal noise.

discriminator method for Oscillators 1 and 2, respectively. The oscillators were then measured against each other using the two-oscillator cross-correlation measurement configuration from Fig. 11. Assuming equal noise in each oscillator, a noise floor of -181 dBc/Hz was obtained (Fig. 15).

Excellent results were obtained with the highly efficient wide tuning range VCO at supply voltages of 9.6 and 16 V [Fig. 8, curves (B) and (C)]. PM noise floors of -180 and -185 dBc/Hz were obtained using the dual-channel frequency discriminator cross-correlation method (Fig. 10). The 1-Hz intercept points were measured as -4 and -1 dBc/Hz, and the output power levels were $+23$ and $+28$ dBm at 9.6 and 16 V supply voltage, respectively. The RF/dc efficiencies were accordingly 28 and 25% for both supply voltages.

The comparison of the phase noise plots for the high- and low-Q STW oscillators [Figs. 14 and 15 versus Fig. 16(A) and (B)] shows the trade-off between the tuning range [Fig. 8, curves (A)–(C)] and the phase noise performance of the investigated STW power oscillators.

Fig. 17 is the residual phase noise of one of the high-Q STW devices characterized in Fig. 1. The value of -142 dBc/Hz for the PM noise at 1-Hz offset is a remarkable result for a 1-GHz metal strip device. The system floor, measured with a 4-dB pad

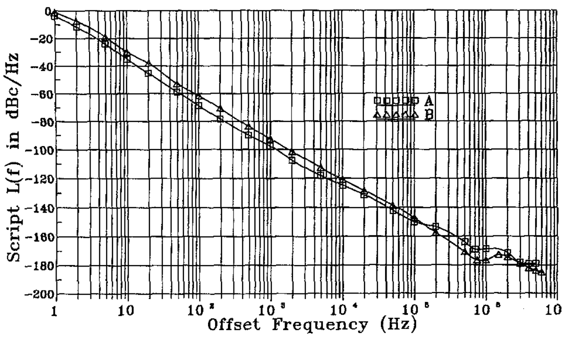

Fig. 16. Phase noise of the wide tuning range VCO at a supply voltage of: (A) 9.6 V (P_{out} = +23 dBm, efficiency = 28%); (B) 16 V (P_{out} = +28 dBm, efficiency = 25%).

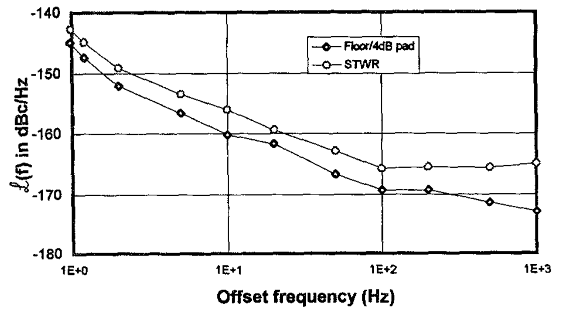

Fig. 17. Residual phase noise of one of the high-Q AQP STW devices with Q_L = 2737 and insertion loss of 3.6 dB.

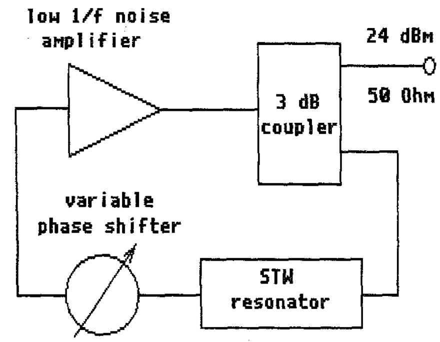

Fig. 18. Test oscillator using a low $1/f$ noise loop amplifier (UTO 1023, Avantek).

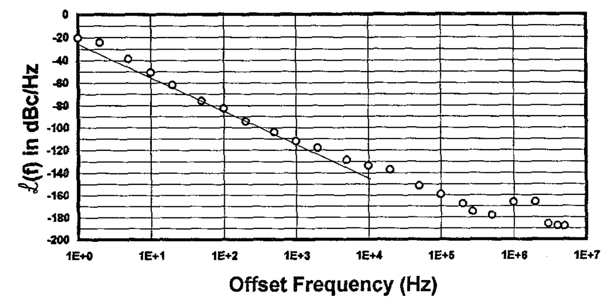

Fig. 19. Phase noise of the test oscillator of Fig. 18 at a gain compression of 7 dB, P_{out} = +24 dBm, P_{loop} = +28 dBm.

instead of the STW device, was 2–4 dB lower than the residual noise of the STW device in the 1–100 Hz range. Both curves show a 10-dB/decade slope for offset frequencies below 100 Hz. The steeper slope below 2 Hz for the STW device and the 4-dB pad was due to noise added by the measurement system. The actual STW device 1-Hz intercept should be approximately −147 dBc/Hz. Further $1/f$ measurements on other high-Q STW devices (AQP STW devices provided by G. K. Montress and T. E. Parker) with a loaded Q of 2600 provided results that were indistinguishable from the system noise floor measured as −142 to −144 dBc/Hz at 1 Hz.

The STW device, characterized in Fig. 17, was incorporated into a simple oscillator loop using a power amplifier which is known for its very low residual noise (UTO 1023, Avantek) [3]. This test oscillator is shown in Fig. 18. It was first run at 7-dB gain compression for minimum loop loss and the output power was +24 dBm. Fig. 19 indicates a PM noise floor of −187 dBc/Hz for this case. The 1-Hz intercept of −21 dBc/Hz was 16 dB worse than what we expected from the residual noise measurement on the STW device (Fig. 17). Also, the slope was −25 dB/decade instead of −30 dB/decade. We found that high gain compression was the reason for this poor performance. After we reduced the gain compression to 1 dB, we obtained the results shown in Fig. 20. The 1-Hz intercept became −33 dBc/Hz, only 4 dB worse than what we would expect if the loop amplifier was ideally noiseless. A measurement with a second low PM noise amplifier from the same manufacturer indicated a 1-Hz intercept of −31 dBc/Hz with the same STW device.

VI. Discussion

The oscillators in Fig. 6 were designed for a phase noise floor of −195 dBc/Hz. We actually measured an 11–14 dB higher noise level. The reason is that these circuits were designed for an STW device loss of 7–8 dB. The devices we used had a loss of only about 3.5 dB. Adding a 4-dB resistive pad to the loop would not solve the problem since resistive attenuators are usually designed for a 50-Ω system. On one hand, this would seriously deteriorate the matching conditions in the loop since AB-class amplifiers have input and output impedances that significantly differ from 50 Ω. On the other hand, the 4-dB pad would lower the drive power level necessary for the AB-class of operation [7]. We tried to increase the insertion loss of the STW device by adding capacitors in series to it, but this also resulted in deterioration of the matching conditions and the amplifier noise figure. It would appear from these results that the effective noise figure of the AB-class amplifier is likely to be very sensitive to the source and load impedances, as well as the extent of gain compression. This problem can be overcome by a second iteration of the amplifier design, which would start after the impedances and insertion loss of the STW devices to be used are known.

The noise floor of the single transistor stage oscillator (Fig. 5) could not be measured because the output power was insufficient for the measurement systems used and there were no identical oscillators available. This concept should also yield very low phase noise floors because it allows high loop power and extremely low loop loss.

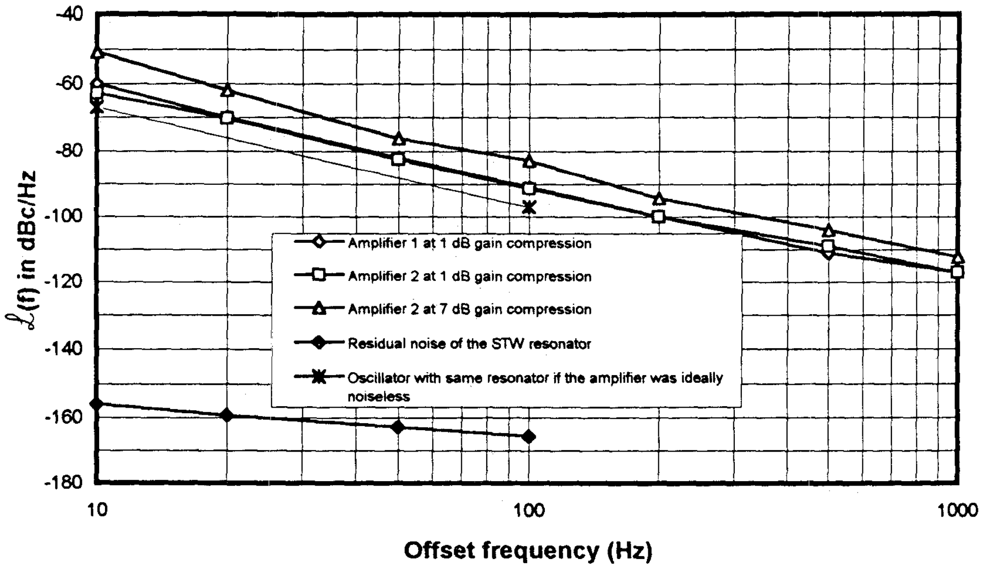

Fig. 20. Phase noise data of STW oscillators using low PM noise amplifiers (UTO 1023, Avantek) and the residual noise of the STW device in these oscillators.

We expected a PM noise floor better than -181 dBc/Hz from the two oscillator cross-correlation measurement (Fig. 5). Unfortunately, strong interference with local AM radio stations caused the bump around 10 MHz and affected the measurement. This bump was not observed with the single-channel frequency discriminator measurement on the same oscillators (see Figs. 9 and 14). This is probably due to the fact that the single channel system was much simpler, easier to set up, and less sensitive to parasitic interference.

VII. Summary and Conclusions

We have demonstrated state-of-the-art 1-GHz STW resonators for high power oscillator applications which feature an insertion loss of 3.6 dB, a loaded-Q of 2740, an unloaded-Q of 8000, and a residual phase noise level of -142 to -144 dBc/Hz at the 1-Hz intercept. Other low-Q resonators and two-pole coupled resonator filters with an insertion loss of 5–9 dB allow tuning ranges of 700–1500 ppm in highly efficient STW power oscillators, which can run from supply and tuning voltage sources below 10 V and are suitable for portable applications. The output power and the noise floor of such VCO's are typically $+23$ dBm and -180 dBc/Hz, respectively. Higher output powers in the $+28$ to $+33$ dBm range and lower noise floors in the -185 to -194 dBc/Hz range can be achieved if higher supply voltages are used (see also [7]). The RF/dc efficiency is typically 25–28%, but a value of 36% was also observed.

One-Watt oscillators running at a loop power of up to $+35$ dBm feature a PM noise floor of -184 dBc/Hz and a 1-Hz intercept of -17 dBm. If a commercially available loop amplifier is used (UTO 1023, Avantek), the phase noise floor can be reduced to -187 dBc/Hz. Such oscillators typically run at a loop power of $+27$ dBm and an output power of $+24$ dBm, and demonstrate a 1-Hz intercept of -21 dBc/Hz. A reduction of the gain compression to 1 dB greatly improves the close-to-carrier phase noise behavior resulting in a 1-Hz intercept of -33 dBc/Hz and a slightly decreased output power of $+22$ dBm.

The results in Fig. 20 represent state-of-the art close-to-carrier phase noise performance of 1-GHz STW high power oscillators. They clearly indicate that the loop amplifier and not the STW device is the major source of $1/f$ noise, even when commercial loop amplifiers with the lowest $1/f$ noise, currently available, are used. Therefore, further work in designing high power loop amplifiers with improved residual phase noise performance is needed.

We believe that STW resonant devices have a strong potential for use in designing extremely low noise, high power oscillators in the lower GHz range. Since STW are surface waves in nature and use very similar device geometries as their SAW counterparts, we feel that STW devices may have the same or similar size dependence for $1/f$ noise as that observed with SAW resonators [15]. If this proves to be true, then increasing the active acoustic area could improve the device residual phase noise to values lower than -150 dBc/Hz at a 1-Hz offset. For a $1/f$ noiseless amplifier, this would correspond to a 1-Hz intercept of -45 dBc/Hz. However, a size dependence for $1/f$ noise in STW devices has not yet been demonstrated.

Finally, using careful loop amplifier design, taking special care for low gain compression, matching, and noise figure, should result in a noise floor below -195 dBc/Hz and a 1-Hz intercept of -45 dBc/Hz in the lower GHz range.

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