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SURFACE TRANSVERSE WAVE OSCILLATORS WITH EXTREMELY LOW THERMAL NOISE FLOORS

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Abstract

This paper presents state-of-the-art results on 1 GHz surface transverse wave (STW) power oscillators running at extremely high loop power levels. 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 phase noise of -142 dBc/Hz at 1 Hz intercept 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 an insertion loss in the 5-9 dB range can conveniently handle power levels in excess of 2 W. These devices were implemented in voltage controlled oscillators (VCO's) running from a 9.6 V source at an output power of 23 dBm and a RF/dc efficiency of 28%. Their tuning range was 750 kHz and the noise floor -180 dBc/Hz. The oscillators, stabilized with the high-Q devices, use specially designed AB-class power amplifiers, deliver an output power of 29 dBm and demonstrate a noise floor of -184 dBc/Hz and a 1 Hz intercept of -17 dBc/Hz. The 1 Hz intercept was improved to -33 dBc/Hz using the UTO-1023 as a loop amplifier. In this case the output power was 22 dBm. In all cases the loop amplifier was the limiting factor for the close-to-carrier oscillator phase noise performance.

1. Introduction

Surface acoustic wave (SAW) based oscillators are well known for their excellent phase noise performance in the frequency range 0.1-1 GHz [1] - [3]. Resonator stabilized oscillators, operating in the 400-500 MHz range and featuring a thermal noise floor of -184 to -185 dBc/Hz were demonstrated recently [3]. Along with a 1 Hz intercept of -48 to -55 dBc/Hz these oscillators are considered to represent the 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 multi-track 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.

This 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 noise floor of -194 dBc/Hz [7]. This oscillator featured excellent tuning and wide band frequency modulation abilities. Its RF/dc efficiency was 19%.

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

2. STW Resonant Devices for Power Oscillator Applications

Compared to SAW, STW offer a greater flexibility in the design of metal strip resonators and narrow band 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 ranging from 500 to 8000 and an insertion loss well below 10 dB at 1 GHz. All these devices can conveniently stand drive power levels in excess of 2 W and were found to operate without measurable performance degradation for several months in different fixed frequency and voltage controlled power oscillators. The design details for such devices have been well documented in references [5], [6], [10] and [11]. Here we will characterize only some of the devices used in this study.

Figures 1 A), B), C) and D) present 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 with all quartz technology [12].

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 it would allow a tuning range of about 180 KHz over the 1 dB device bandwidth as evident from Fig. 1 B).

A much wider tuning range can be achieved with the low-Q resonator characterized in Fig. 2. It has an insertion

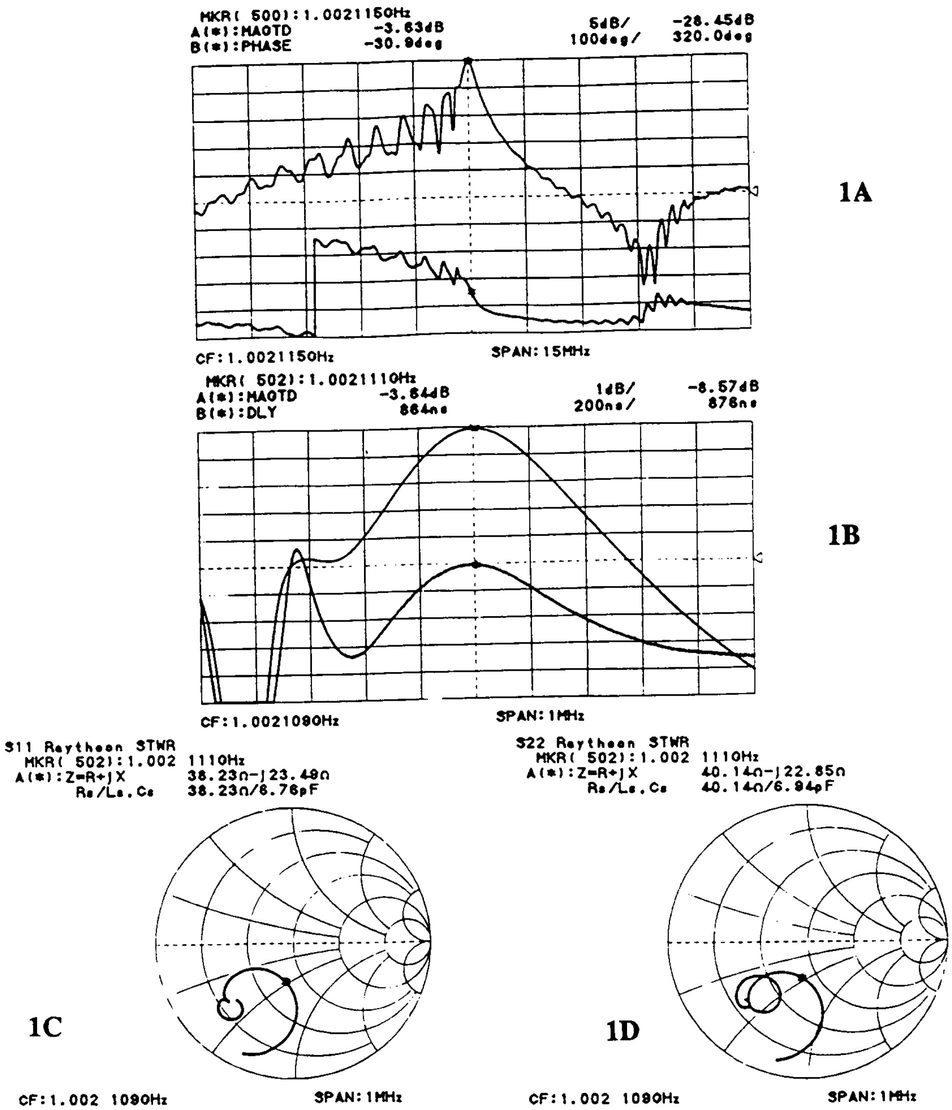

Fig. 1. Characteristics of a single mode high-Q device realized at Raytheon:
A) Frequency and phase responses,
B) Detailed frequency and group delay responses,
C) Input reflection coefficient, S11,
D) Output reflection coefficient, S22.

loss of 5.2 dB, a loaded Q of 1500 and allows a tuning range of about 700 KHz if the VCO is tuned over the 3 dB device bandwidth.

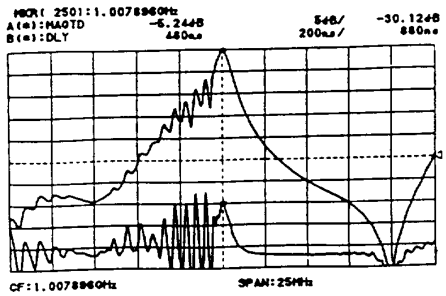

Fig. 2. Frequency and group delay responses of an STW resonator with $Q_l = 1500$ and 5.2 dB insertion loss.

If even wider tuning ranges are necessary, the 2-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 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 cascading 3 C-L-C varactor tuned phase shifters would be necessary [4].

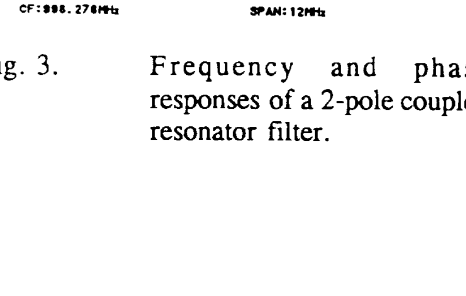

Fig. 3. Frequency and phase responses of a 2-pole coupled resonator filter.

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.

2. 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 2 W. 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. Even if some power transistors can generate high output power in the A-class of operation and a 50 Ω environment, their efficiency rarely exceeds 10%. An elegant and inexpensive solution of 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 B- or C-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 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 and D). Under these circumstances measuring the oscillator loop power becomes a serious problem. Just breaking the loop 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 series 0.47 pF capacitor soldered to the central line.

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 strip lines and ground planes, surrounding them, as shown in Fig. 4. The reading is obtained by a spectrum analyzer or power meter connected to the other end of the cable. The probe is calibrated by touching the load at the oscillator output at which 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 oscillator tuning range in which the output amplitude was constant.

We investigated three types of high loop power STW oscillators using highly efficient AB-class loop amplifiers. Since the operation principle of such oscillators was described in detail in [7], we 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.

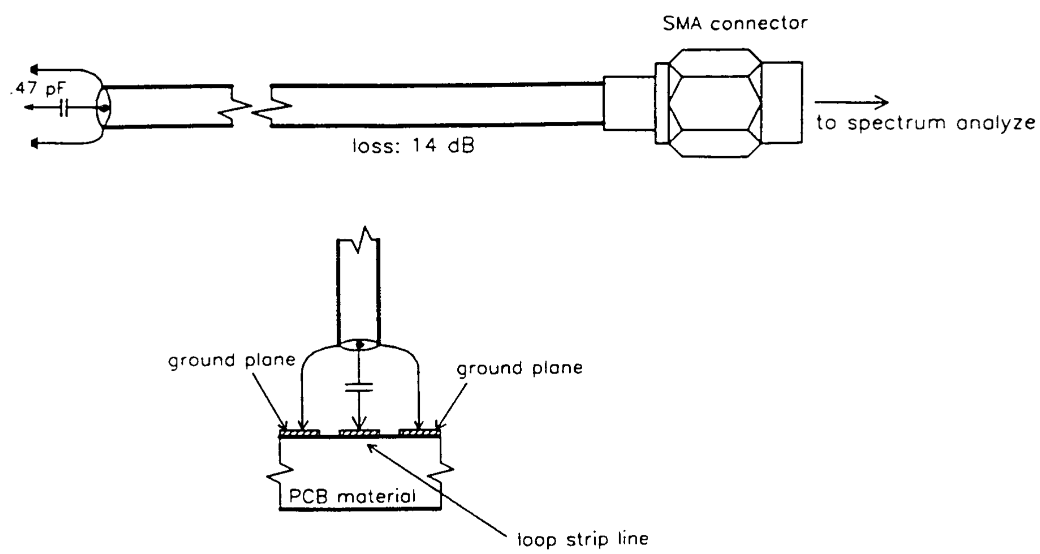

Fig. 4. A capacitive probe for evaluation of the oscillator loop power under closed loop conditions.

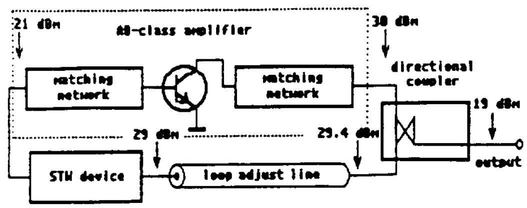

Fig. 5 Simple STW high loop power oscillator with directional coupling to the load.

This was achieved using a 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 with respect to load changes. Unfortunately, we were unable to achieve usable frequency tuning with this design. The addition of a varactor tuned phase shifter in the loop was found to make the oscillator unstable with tuning due to a deterioration of the matching condition. Another drawback of this design was that the gain compression had to be kept very low (1-2 dB) which was necessary because an AB-class amplifier is a poor limiter. Increasing the loop power results in an increase of the collector current which can thermally overload the transistor.

The tuning and limiting problem could readily be 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 two-fold. First, it provides sufficient isolation between the VPS and the power stage and second, it performs the limiting function providing a safe input power level to the power transistor. Thus excellent tuning over the 3 dB device bandwidth and

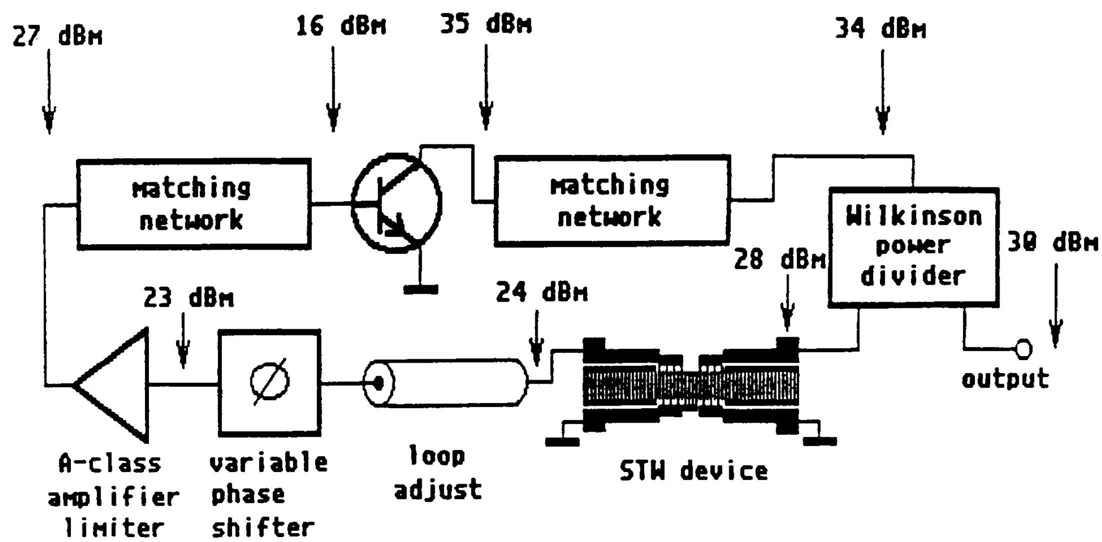

Fig. 6. Block and level diagram of a 1 W VCO stabilized with the high Q device from Fig. 1.

stable amplitude over the entire tuning range are guaranteed. 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 shown in Fig. 19 below. A trade-off between gain compression, loop loss, device Q and tuning range can be achieved by adding series capacitors to the STW device. 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 noise floor of -195 dBc/Hz [3].

Figure 7 shows a highly efficient 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 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%.

Curves 8 A, B and C in Fig. 8 show the tuning characteristics of the power VCOs from fig.s 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

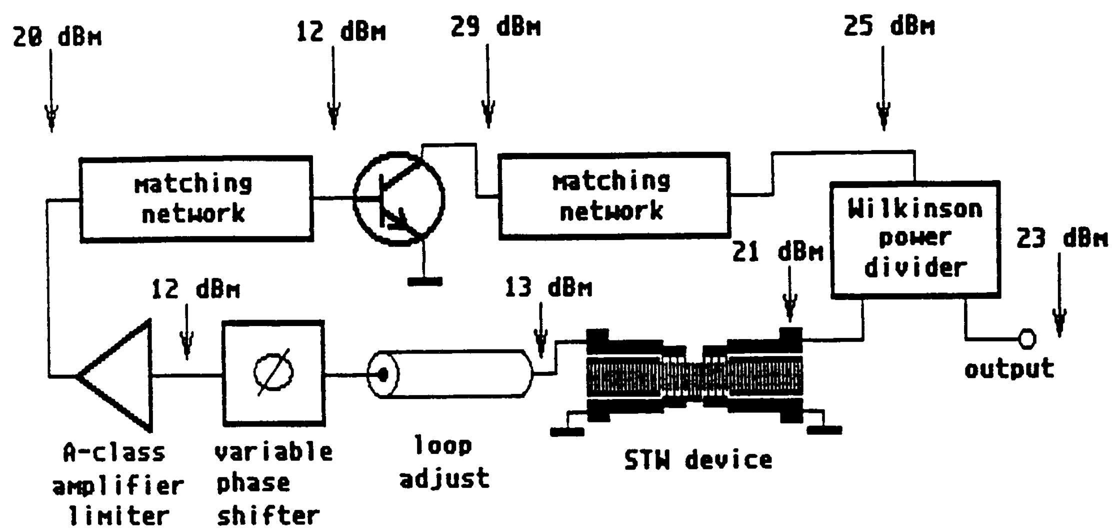

Fig. 7. 9.6 V VCO for portable applications with RF to dc efficiency of 28%.

phase shifters and using the CRF from Fig. 3 would increase the tuning range to 1.5 MHz with a tuning voltage of 0 to 9.5 V.

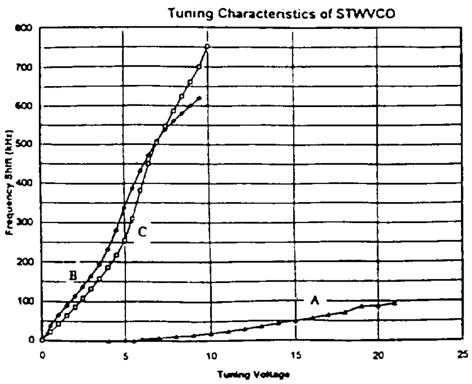

Fig. 8. Tuning characteristics of the power VCO:
A) 1W VCO using the high-Q device from Fig. 6,
B) 9.5 V wide tuning range VCO from Fig. 16 B,
C) Wide tuning range VCO at 16 V supply voltage from Fig. 16 C.

3. 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 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 power oscillators to make sure that the systems delivered comparable results.

First we tried to measure the oscillator noise floor. The simplest and most forgiving system for noise floor evaluation is the single channel frequency discriminator with coaxial cable of delay τ_d 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 $(f < \frac{0.35}{\tau_d})$ [13]. Moreover, it needs only one oscillator and adapts to small changes of the oscillator frequency during the measurement.

If the outputs of two identical channels of this system are cross correlated in a 2-channel FFT analyzer (Fig. 10) then the uncorrelated system noise cancels out and the system sensitivity can be improved by 20-25 dB [13] - [15].

This cross correlation concept can also be applied to a measurement system using two identical oscillators (Fig. 11). The system noise floor in this case can exceed -195 dBc/Hz even at very high Fourier frequencies [13] - [15].

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Residual phase noise measurements on the high-Q STW device from Fig. 1 were

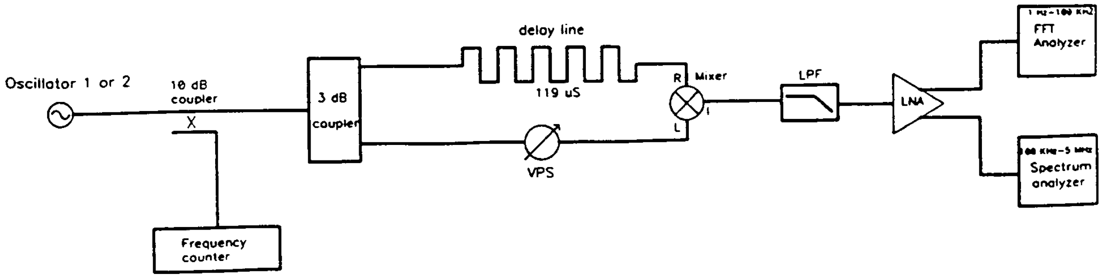


Fig. 9. Measurement setup using a single channel delay line frequency discriminator.

performed with the setup in Fig. 12, adapted from reference [4]. A low PM noise frequency synthesizer [16] was used to make phase noise measurements up to 1kHz 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 [17], 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 [17] were performed with the system setup in Fig. 13. This setup assumes that the reference source (in this case a low-noise frequency synthesizer) is substantially quieter than the STW oscillator. In our case this condition breaks down for offsets greater than 1 KHz.

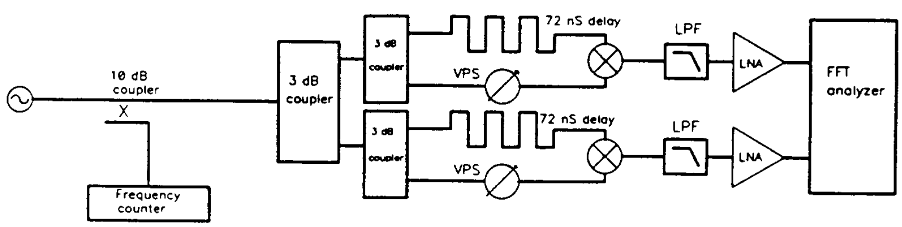


Fig. 10. Cross-correlation frequency discriminator measurement setup.

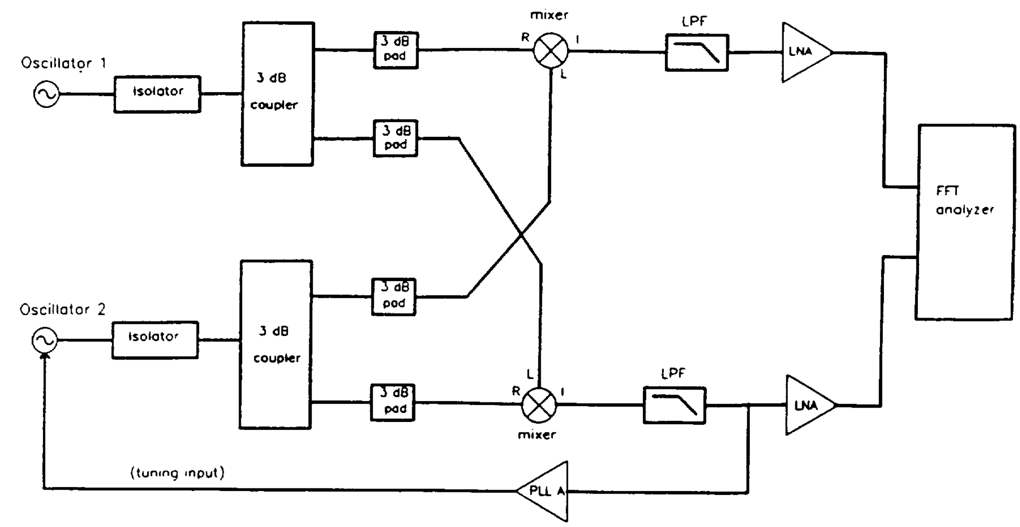


Fig. 11. Two oscillator cross-correlation measurement.

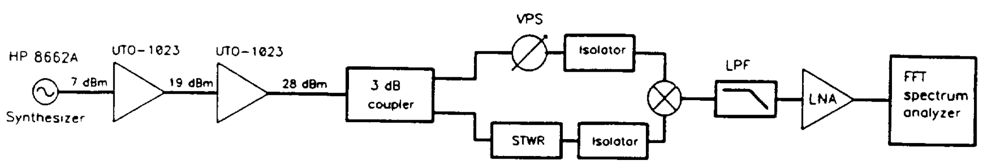


Fig. 12. Setup for residual phase noise measurements on high-Q STW devices.

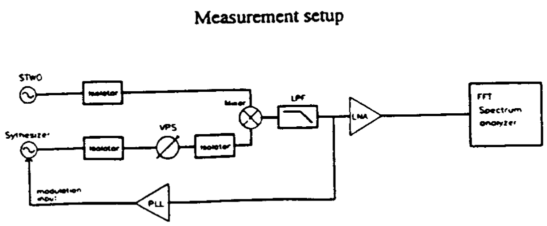


Fig. 13 System for measuring the close-to-carrier phase noise of fixed frequency STW oscillators.

4. Evaluation of the Phase Noise Data

Figure 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 was in this case 29 dBm. A noise floor of -184 and -182 dBc/Hz was measured with the single channel frequency discriminator method for Oscillator 1 and Oscillator 2 respectively. Then both oscillators were 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).

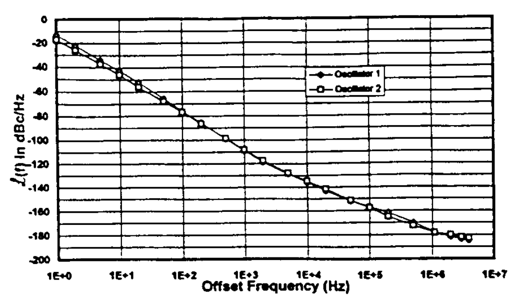


Fig. 14 Phase noise plots of two high-Q STW power oscillators measured with the single channel frequency discriminator method. Both oscillators have $Q_l=4000$, $P_{out}=29$ dBm.

Excellent results were obtained with the highly efficient wide tuning range VCO at supply voltages of 9.5 and 16 V (Curves 16 A and B in Figure 10). 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 9.5 and 16 V supply voltage respectively. The RF/dc efficiencies were accordingly 28% and 25% for both supply voltages.

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

Figure 17 is the residual phase noise of one of the high-Q STW devices characterized in Fig. 1. The value of -142 dBc/Hz at 1 Hz intercept is a remarkable result for a 1 GHz metal strip device. The system floor, measured with a 4 dB pad 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 caused by the measurement system. The actual STW device 1 Hz intercept is approximately -147 dBc/Hz.

This same STW device was implemented in an a simple oscillator loop using a power amplifier which is known for its very low residual noise [3], [17]. This test oscillator is shown in Fig. 18. First it was run at 7 dB gain compression for minimum loop loss. The output power was

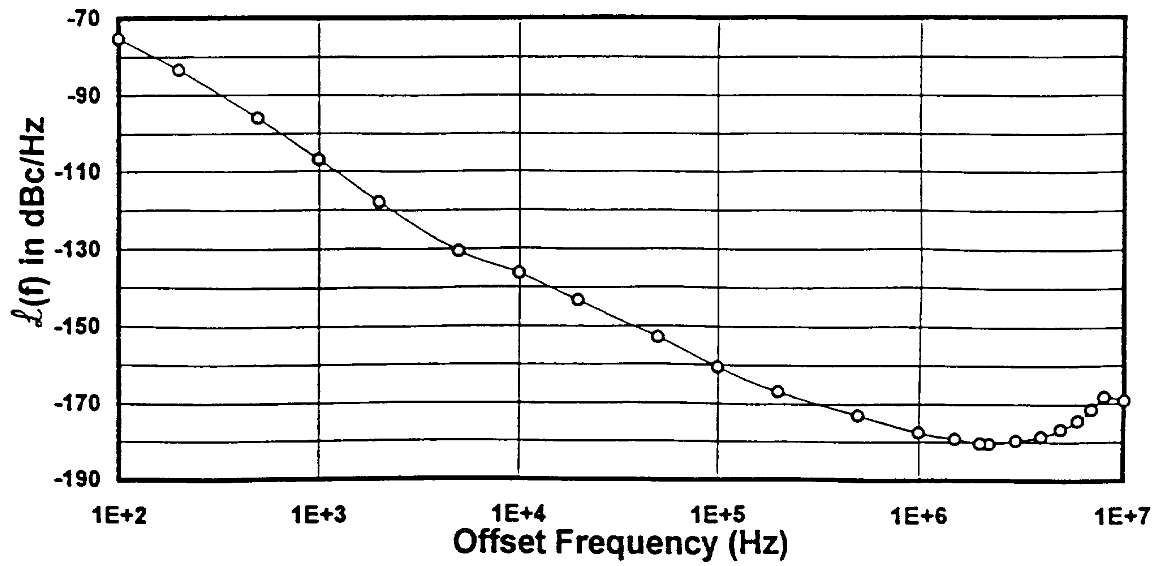


Fig. 15. Phase noise data of the high-Q oscillators obtained from the two-oscillator cross correlation measurement (Fig. 11) assuming equal noise.

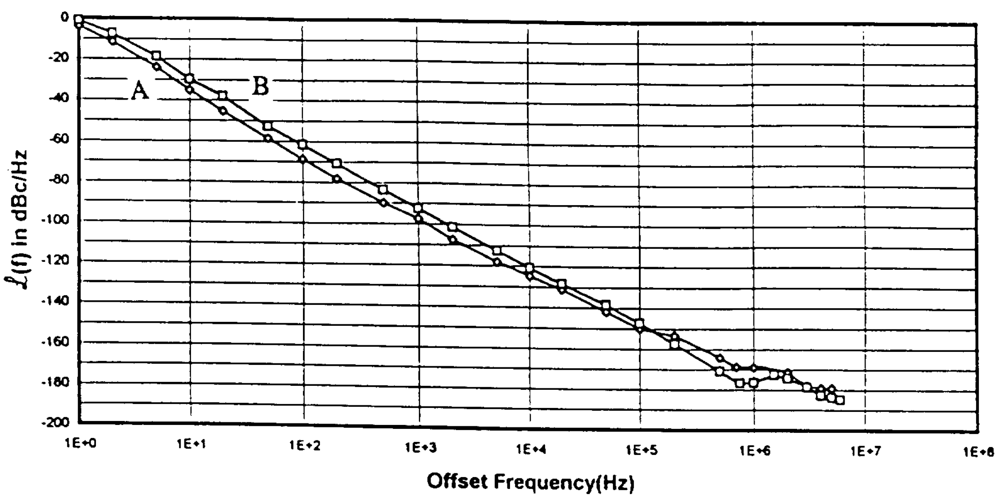


Fig. 16. Phase noise of the wide tuning range VCO at a supply voltage of:
A) 9.5 V (P_{out}=23 dBm, efficiency=28%) and
B) 16 V (P_{out}=28 dBm, efficiency=25%).

24 dBm. Figure 19 indicates a 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

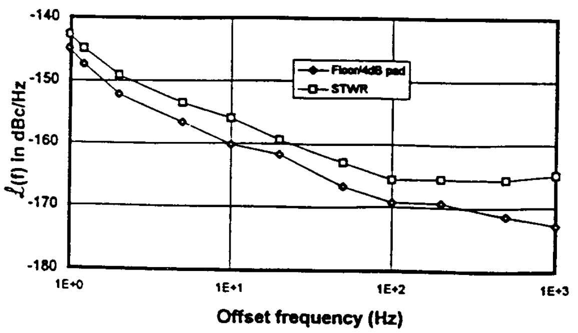


Fig. 17. Residual phase noise of one of the high-Q STW devices fabricated at Raytheon with Q_L=2737 and insertion loss of 3.6 dB.

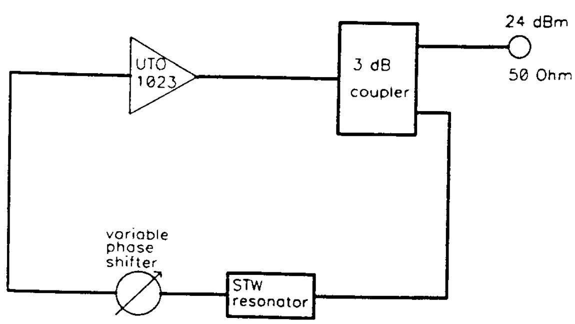


Fig. 18 Test oscillator using a low 1/f noise loop amplifier.

instead of 30 dB/decade. We found that the high gain compression was the reason for this poor performance. After we reduced the gain compression to 1 dB we obtained the results in Fig. 20. The 1 Hz intercept was -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 amplifier from the same manufacturer indicated a 1 Hz intercept of -31 dBc/Hz with the same STW device [17].

5. Discussion.

The oscillators in Fig. 6 were designed for a noise floor of -195 dBc/Hz. We measured 11-14 dB less. 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 about 3.5 dB. We increased this loss by increasing the loaded Q but this resulted in deterioration of the matching conditions and the amplifier noise figure. We believe that this problem can be overcome by more careful amplifier design which should start after one knows the impedances and insertion loss of the STW devices to be used.

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

We expected 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 Fig. 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.

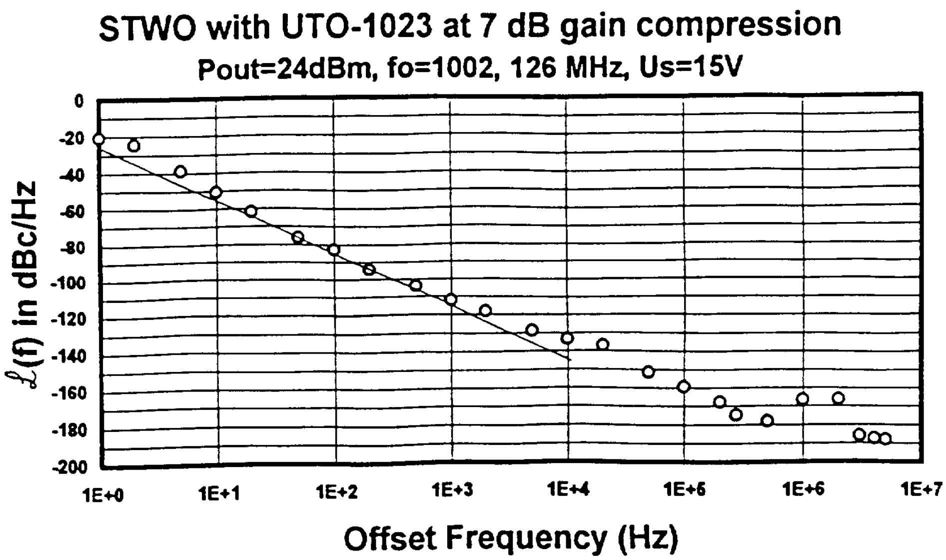
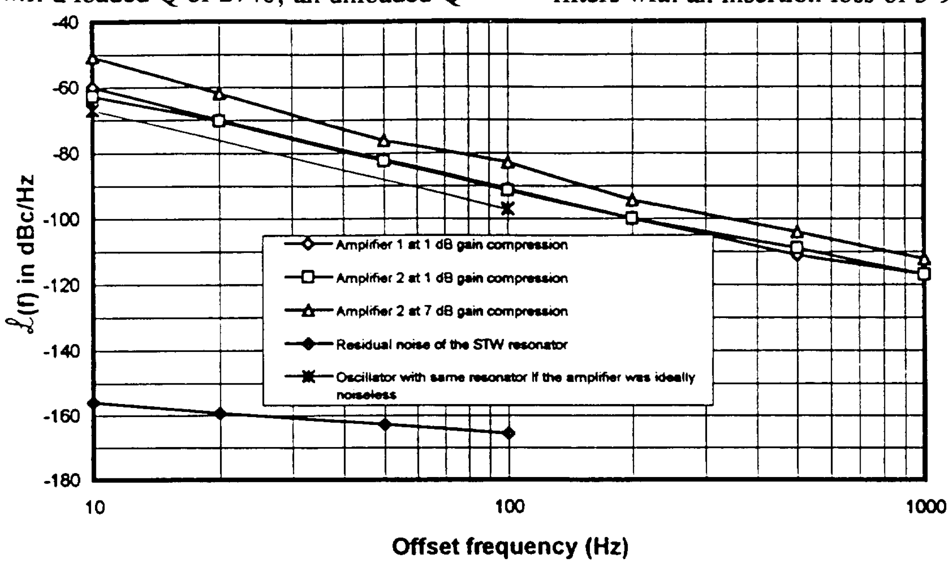


Fig. 19. Phase noise of the test oscillator of Fig. 18 at a gain compression of 7 dB, P_{out}=24 dBm and P_{loop}=28 dBm.

6. Summary and conclusions.

We have demonstrated state-of-the-art 1 GHz STW resonators for power oscillator applications featuring an insertion loss of 3.6 dB. a loaded Q of 2740, an unloaded Q of 8000 and a residual noise level of -142 dBc/Hz at 1 Hz intercept. Other low-Q resonators and two-pole coupled resonator filters with an insertion loss of 5-9 dB allow



Fig

Fig. 20. Phase noise data of STW oscillators and the residual noise of the STW device with low noise loop amplifier [18].

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 are typically 23 dBm and -180 dBc/Hz respectively. Higher output powers in the 28-33 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 Reference [7]). The RF/dc efficiency is typically 25-28% but a value of 36% was also observed.

1 W oscillators running at a loop power of up to 35 dBm feature a noise floor of -184 dBc/Hz and a 1 Hz intercept of -17 dBm. If a low 1/f noise loop amplifier is used the noise floor can be reduced to -187 dBc/Hz [17]. Such oscillators typically run at a loop power of 28 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 the state-of-the-art close-to-carrier phase noise performance of 1 GHz STW power oscillators. They clearly indicate that the loop amplifier and not the STW device is the major source of 1/f noise even if the best loop amplifiers currently available in the market are used. Therefore, further care has to be taken in designing high power loop amplifiers with improved residual phase noise performance.

We believe that STW resonant devices have a strong potential for use in designing extremely low noise power oscillators in the lower GHz range. Increasing the active acoustic area could improve the device residual phase noise to values lower than -150 dBc/Hz at 1 Hz intercept. Using careful loop amplifier design and taking special care of low gain compression, matching and noise figure will result in noise floor values below -195 dBc/Hz and a 1 Hz intercept of -45 dBc/Hz in the lower GHz range.

Acknowledgements

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