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THE RF-POWERED SURFACE WAVE SENSOR OSCILLATOR - A SUCCESSFUL ALTERNATIVE TO PASSIVE WIRELESS SENSING

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Abstract - **This paper describes a novel passive wireless SAW sensor providing a highly coherent measurand proportional frequency, FM modulated with identification (ID) data and immune to interference with multiple path signals. The sensor comprises a highly stable low-power oscillator, stabilized with the sensing SAW resonator and powered by the rectified RF power of the interrogating signal which is received by a λ/4 whip antenna on the sensor part. A few hundred µW of dc power are enough to power the sensor oscillator and ID modulation circuit and achieve stable operation at 1.0 and 2.49 GHz. Reliable sensor interrogation was achieved over a distance of 0.45 m from a SAW based interrogation unit providing 50 mW of continuous RF power at 915 MHz. The -30 to -35 dBm of returned sensor power was enough to receive the sensor signal over a large distance and through several walls with a simple superheterodyne FM receiver converting the sensor signal to a low measurand proportional intermediate frequency and retrieving the ID data through FM detection. Different sensor implementations including continuous and pulsed power versions and the possibility of transmitting data from several measurands with a single sensor are discussed.**

I. INTRODUCTION

Since the early years of radio broadcasting it has been known that radio waves can transmit electrical power over a certain distance. People living in the vicinity of AM-radio transmitter antennas have been using this feature to power electrical bulbs, charge batteries, etc.. One of the most promising modern applications of power transfer with radio waves is passive wireless remote sensing using surface acoustic wave (SAW) devices [1], [2]. The sensing device which is a SAW-based tapped or phase coded delay line, single-port resonator or correlat.ponse of the SAW sensor. The interrogation pulse, also called request signal [2], is received by a small antenna on the sensor part and, after the acoustic time delay, the sensor response, carrying information about temperature, pressure, torque, acceleration, gas concentration and other physical quantities at the sensor location, including sensor ID information, is transmitted back through the same antenna for reception and data processing by a remotely located receiver unit [1-4]. Although, in most cases, passive sensor devices are inexpensive if fabricated on a large scale, and the principle of wireless SAW sensing is fairly simple, it has several drawbacks. For efficient excitation, the frequencies of the interrogation burst and sensor device must be identical over a narrow tolerance. This problem is typically solved by using swept RF-bursts but this increases circuit complexity on the transmitter part. If the sensor has to transmit also ID information, then fabrication of unique and complex sensor devices is necessary. One of the most severe problems is the interference of the sensor signal with multiple path signals, launched by the interrogation pulse, and antenna mismatch which strongly depends on the properties of the RF-propagation channel [1], [4]. This may greatly degrade the received sensor response and reduce reliability in obtaining correct sensor data. This problem is again solved with more complex differential sensor devices [4] and increased circuit and data processing complexity on part of the radar-like transceiver [2]. In some applications, error free data acquisition is obtained by greatly limiting the distance between sensor and interrogation unit [3]. All these drawbacks become very severe in the frequency range of 2.4 to 2.5 GHz which has been allocated as a world standard for SAW-based RF sensors [1]. At such high frequencies, due to various loss mechanisms in the SAW device, the sensor efficiency is fairly low and antenna mismatch and multiple path interference become difficult to handle.

This paper suggests a different approach to passive wireless SAW sensing which allows much more reliable generation, propagation and detection of the sensor response. Instead of exciting the SAW sensor directly, the power of the interrogating RF-burst is converted into dc power and used to activate a microwave oscillator, which is stabilized with the sensor device, taking advantage of its high Q and low loss. Once the sensor oscillator (SO) starts, it generates a RF signal with a highly coherent frequency which depends to a large extent on the measurand and very little on the properties of the request signal or its RF-propagation channel. Also, the SO frequency is immune to interference with multiple-path or other stray signals. In addition, the sensor can be FM, AM, FSK or PM modulated by a low-power CMOS circuit, also dc supplied from the interrogation pulse power, to transmit ID information and data from several measurands using a single SO. This greatly increases sensor design flexibility and functionality. Moreover, since the interrogation signal is converted into dc, its frequency does not need to be identical with the SO frequency.

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In this way, the request unit and SO can operate at two different frequencies within the same frequency band or even in two different frequency bands allocated for wireless sensor applications. On one hand, this provides total separation of the interrogation and SO signals. On the other hand, the design of the request and receiver units and processing of the sensor data and ID information are greatly simplified. Finally, the transmitted SO signal can be received over a large distance and without line-in-sight conditions using a simple superheterodyne receiver, down converting its frequency into a convenient to process, measurand proportional low intermediate frequency (IF) and retrieving ID and modulation data through simple AM or FM detection. All this increases reliability, reduces error probability and allows the design of simple portable systems for quick data acquisition from several sensors. The sensor system can be designed to operate on continuous interrogating power for instant sensor response or for pulsed power, accumulated in a capacitor at the sensor location and powering the oscillator after a certain integration time. The pulsed power method is useful if a larger distance has to be covered with a limited interrogation power level.

II. Operation Principle

The SO operation principle is illustrated in Fig. 1 and based on converting the energy of the interrogating RF signal with the frequency F1 into dc power which is then converted into RF power at another frequency F2, identical to that of the of the SAW sensor device in the moment of the measurement. Thus the SO acts as a RF-powered RF transmitter generating a highly coherent measurand proportional frequency F2. If the request unit provides a continuous wave (CW) signal, after reception by the Rx antenna and RF detection on the SO part, it will be converted into a dc supply voltage (Us) proportional to the received excitation power. This dc voltage is applied to the power supply terminal of the feedback-loop oscillator (FLO), stabilized with a high-Q sensing SAW or surface transverse wave (STW) two-port resonator (TPR). In a properly designed SO, the generated frequency will track exactly the measurand proportional resonator center frequency which is then transmitted by the Tx antenna and can be remotely received and measured to extract measurand specific information at the sensor location. The power splitter in Fig. 1 provides optimum matching of the SO power to the Tx antenna and isolates its variable impedance from the oscillator loop.

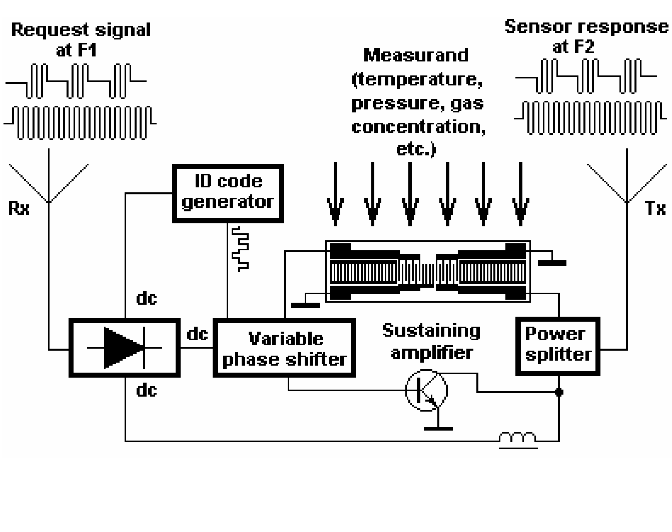


Fig. 1. Block diagram of a typical FLO based sensor oscillator.

If the request signal is a RF burst and its pulse duration is larger than the SO start-up time, then the SO signal will also be a RF burst sequence, whose carrier frequency will be proportional to the effect of the measurand.

In Fig. 1, the dc voltage from the request signal is also used to power an ID code generator and provide bias to a varactor controlled variable phase shifter (VPS) to allow FM or FSK modulation of the SO frequency to transmit ID data and/or data from one or more additional measurands extracted from other nonacoustic sensors at the SO location. If, for example, the surface wave resonator is mounted on a thin membrane to measure pressure, the SO carrier frequency will be pressure proportional, while information on the temperature at the sensor location can be provided by the low frequency of a thermistor controlled RC oscillator driving the ID code generator whose output FSK modulates the SO carrier. Thus the SO will carry information on pressure and temperature and will identify itself.

In some countries and frequency bands the maximum transmitted power for wireless sensor applications is very limited, e. g. in Germany, it is just 10 dBm in the 433 MHz band. This will limit the SO interrogation range to less than 0.5 m. If the sensor system has to bridge a larger distance with such a low request power, then the SO circuit from Fig. 2 is more appropriate. The request signal is a train of RF bursts. After RF detection by D1, the carrier is removed and the amplitude of the low-frequency pulses is multiplied by a desired factor using the step-up transformer Tr1. The second detector D2 rectifies the stepped-up pulse signal, converting it to a dc voltage with an amplitude much higher than the peak level of the incoming RF signal. Since the output impedance of Tr1 is fairly high, the rectified dc voltage cannot drive the SO directly, however, it can charge the capacitor C. Once the charge voltage reaches a certain threshold level, applied to the enable (en) pin of the voltage regulator through the high-impedance resistors R1 and R2, the voltage regulator enables a discharge of the stored in the

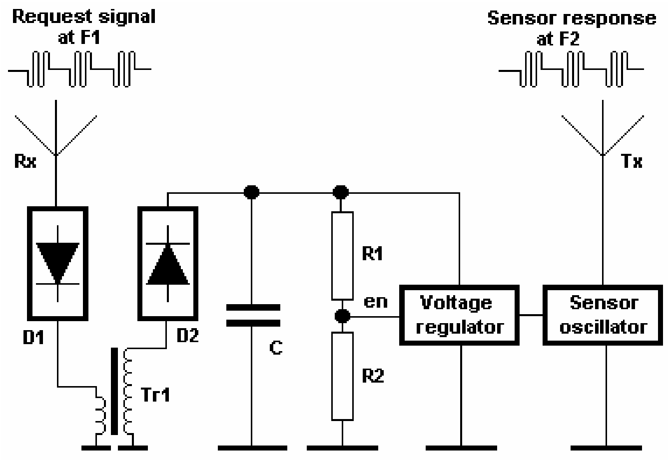


Fig. 2. SO operating on stored energy obtained from the request RF bust.

capacitor C dc energy through the SO which starts transmitting sensor information until the capacitor is discharged below the oscillation level. Then the voltage regulator is disabled and the charging cycle starts all over again. The SO response time is equal to the capacitor charge time which depends on the duration of the incoming request signal and its level at the Rx antenna. Practically, due to losses in the step-up transformer and voltage regulator, the practical SO response time is limited to about 1s at the largest distance to the request unit over which the SO still responds to a given request power. Current studies on SO, according to Fig. 2, are on the way.

III. Design Aspects of RF-powered Sensor Oscillators

The main problem in designing a highly stable RF-powered sensor oscillator is to minimize the dc power necessary for oscillation to start. Another problem comes from the limited amount of interrogation power. As shown further in this paper, very small part of that power reaches the reception Rx antenna of the SO unit. Because of that, the dc supply voltage for the SO rarely exceeds 2V even at fairly small distances to the request unit. These limitations require practical SO to oscillate at dc supply voltages well below 1V and draw a current of no more than 1 to 2 mA. The lower the SO start-up dc power, the larger the interrogation distance at a given amount of RF power. The third problem comes from the fact that the SO supply voltage is directly proportional to the level of the received RF signal and varies over a wide range dependent on the propagation properties of the RF channel. This requires a high stability of the SO frequency with respect to supply voltage variations to minimize sensor frequency error coming from these variations. Finally, the SO frequency has to be sufficiently stable with respect to load variations since the impedance of the transmitting Tx antenna will also vary with the distance to conducting objects within the RF propagation channel.

Fortunately, recent advances in low-loss and high-Q SAW and STW based resonant devices at frequencies up to 3 GHz and the availability of low-cost semiconductors, providing high-gain at microwave frequencies, allow all these problems to be overcome with fairly simple, low-cost SO circuits.

A. Semiconductor Devices for the SO Sustaining Amplifier

A variety of low-power, high-gain microwave transistors, suitable for SO applications are available at a low cost from different manufacturers. Table 1 summarizes specification

TABLE 1
High-gain Microwave Transistors for SO Applications

Transistor Parameter	NE 68618 NEC	MT3S31T Toshiba
Transition frequency Ft	15 GHz	19 GHz
Power dissipation Ptot	30 mW	100 mW
Bias conditions	Uce=0.5V/Ic=0.5mA	Uce=3V/Ic=10 mA
Power gain (MAG) at:		
433 MHz	13.9 dB	no data available
915 MHz	10.5 dB	19 dB
2500 MHz	6.9 dB	11 dB

data from two Si-bipolar transistor models that worked well in the SO subject of this work. The NE68618 features a transition frequency (Ft) of 15 GHz and needs a collector voltage Uce=0.5V and a collector current Ic=0.5mA, which is only 250 µW of dc power, to provide almost 7 dB of maximum available gain (MAG) at 2.5 GHz. The other MT3S31T model was found to provide similar performance at low dc power, although 30 mW are necessary to achieve the maximum gain of 11 dB. If such transistors are used as sustaining amplifiers in the 2.5 GHz SO according to Fig. 1, then 5 to 6 dB of loop loss will provide stable oscillation at that frequency. At the other two frequencies: 433 MHz and 915 MHz, also used for wireless applications, the gain is much higher, therefore, SO operation with higher loop loss is possible, e.g. in SAW sensors coated with a chemosensitive layer, causing higher sensor loss [5].

B. Surface Wave Resonant Devices Appropriate for Sensor Oscillator Applications

Recent developments in SAW/STW sensor technology have shown that TPR and inline coupled resonator filter (ICRF) devices on quartz are an excellent choice for a variety of sensor applications, since they feature very low insertion loss, high phase slope, excellent temperature stability and can stand substantial loading by additional sensing layers [5], [6]. For these reasons, TPR and ICRF are the logical choice for the proposed SO. The 1 GHz STW based TPR device, characterized in Fig. 3 and used in one of the SO designs from Fig. 1., has an unmatched insertion loss of only 4.2 dB. The power splitter and variable phase shifter add another 4.5 dB loss, increasing the overall loop loss to 8.7 dB. Both transistors from Table 1 were found to readily compensate for this loss and provide stable SO operation at less than 0.2 mW of dc power. Another important feature of this device is its moderately high loaded Q of about 3000 which allows linear FM or FSK with a 150 KHz modulation bandwidth. This bandwidth is sufficient for the SO to transmit ID information and data on several measurands from nonacoustic sensors at the SO location.

Even lower insertion loss and higher phase slope is possible with ICRF devices operating up to 5 GHz using the STW mode [7]. The 2.488 GHz device from Fig. 4, described in [7] in detail, has an insertion loss of only 3.5 dB and was found to provide stable SO oscillation at only about 0.5 mW of dc power. The large 2 MHz 1 dB device bandwith allows broadband FM and FSK and can be used to accommodate huge amounts of sensor data generated by low-power CMOS circuitry at the SO location. ICRF devices in the 2.4 to 2.5 GHz frequency range, as the one from Fig. 4, are expected to become available at a low cost in the near future [8].

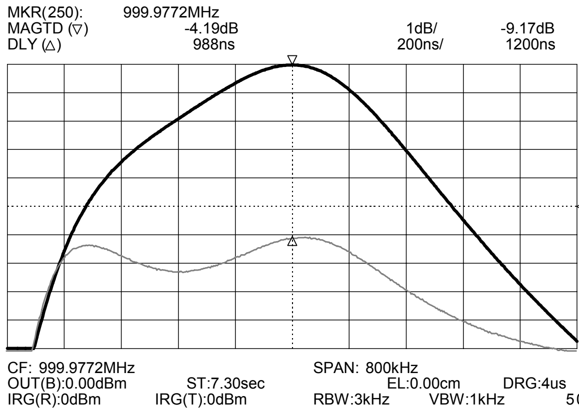


Fig. 3. Frequency (upper curve) and group delay (lower curve) responses of a 1 GHz STW based TPR used in a FSK modulated SO.

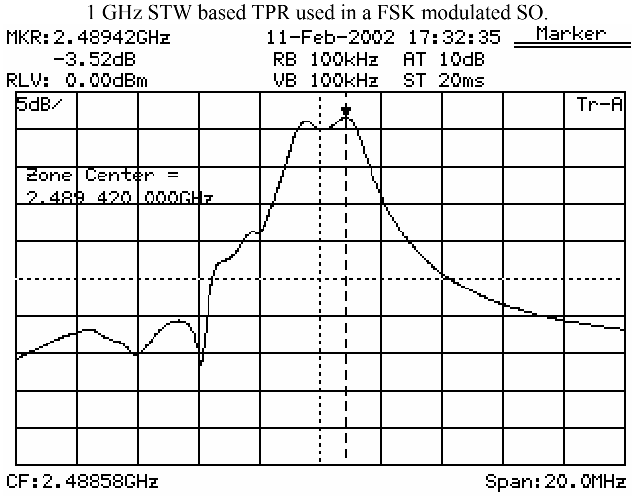


Fig. 4. Frequency response of a 2.488 GHz STW based ICRF well suited to SO applications in the 2.4 to 2.5 GHz range.

Another approach to realizing the proposed SO is by designing it as a negative resistance oscillator (NRO) [9], [10], stabilized with a single-port resonator (SPR) as the one characterized in Fig. 5. NRO take advantage of the very high loaded Q and low motional resistance achievable with SPR devices in the lower GHz range [9]. This reduces SO circuit complexity, allows stable SO excitation at even lower dc power levels and greatly improves the SO frequency stability w.r.t. supply voltage and antenna impedance variations. The device from Fig. 5 has a Q of about 8000 and a motional resistance value of 12Ω which results in a very low 1 dB insertion loss at series resonance (marker position in Fig. 5). Despite the high device Q, NRO allow several 100 KHz of tuning range [9]. The SPR phase response in Fig. 5 is far from linear, therefore, tuning linearity is not as good as in TPR stabilized FLO, but wideband FSK operation is still possible. Within this study, NRO based SO, stabilized with 1.247 GHz and 915 MHz SPR devices as the one in Fig. 5, were found to start at Us=0.6 to 0.7Vdc and provide stable operation at less than 200 µA of dc current.

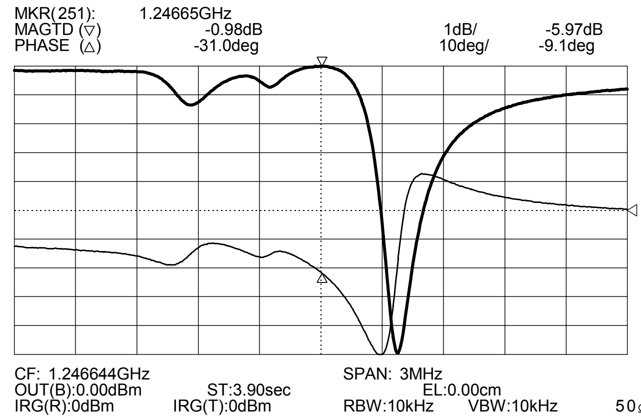


Fig. 5. Frequency (upper curve) and phase (lower curve) responses of a 1.247 GHz SPR obtained in a Pi-circuit measurement.

IV. EXPERIMENTS

Within this study, the operation principle of the proposed SO principle was verified on FLO based SO circuits according to Fig. 1 and optimized for operation at 1 GHz and 2.49 GHz, using the devices from Fig. 3 and 4, respectively. Anoter 2.49 GHz NRO based SO was stabilized with the 1.247 GHz device from Fig. 5 and evaluated at its 2.49 GHz second overtone. Its design was similar to the NRO designs presented in [9] and [10]. The FLO sensors were interrogated at 915 MHz by a portable interrogation unit containing a battery operated STW oscillator and generating 17 dBm (50 mW) of continuous or pulsed interrogation power. The 2.49 GHz NRO based SO was interrogated at the same 915 MHz frequency, however, the request power was increased to 27 dBm to excite a 2-nd overtone strong enough for remote evaluation.

A. The FLO Based Sensor Oscillators

The electrical performance of the two FLO based SO is given in Table 2 . The consumed dc power and the RF output power were measured statically, by connecting a variable dc power supply to the Us terminal and increasing Us until stable oscillation is established. In the distance measurement, two λ/4 whip antennas, Rx at 915 MHz and Tx at the SO frequency, were connected to the stand-alone SO, as shown in Fig. 1, and the interrogation unit was approached to the sensor until oscillation starts. At this position, the distance between SO and interrogation unit was recorded. The transmission of ID data was simulated by an audio frequency low-power flip flop, connected to the variable phase shifter, and FSK modulating the SO frequency with a 400 Hz square wave signal. The flip flop was found to consume less than 10% of the available dc power provided by the request RF signal. The SO signal was received and evaluated by a remotely located double conversion superheterodyne receiver and a spectrum analyzer with an antenna, connected to its input.

The power *Pr*, received by the Rx antenna of the SO unit, is a function of the interrogation wavelength λ, the transmitted request power *Pt*, the gains *Gt* and *Gr* of the transmission and reception antennas of interrogation and SO units, respectively, and the distance *R* to the request unit, as follows [11]:

$$Pr/Pt = (1/4\pi)^2 GtGr(\lambda/R)^2 \tag{1}$$

With *Gt, Gr*=1 (0 dB), *Pt*=50 mW and λ = 32.8 cm at 915 MHz, the received RF-power *Pr* will be about 270 µW if the SO is interrogated from a distance *R*=0.5 m. The 270 µW *Pr* value, calculated with (1), is in a fairly good agreement with

TABLE 2
ELECTRICAL PERFORMANCE OF SO UNITS INTERROGATED AT 915 MHZ WITH 17 dBm OF RF POWER

Sensor oscillator	1 GHz FLO	2.49 GHz FLO
Supply voltage Us	0.8 V	1.015V
Supply current (Is)	200 µA	500 µA
Consumed dc power	160 µW	508 µW
Output power (Pout)	-30 dBm	-35 dBm
Maximum. request range (R)	45 cm	20 cm

the dc power of 160 µW necessary for the 1 GHz SO to start (see Table 2). The reason why *Pr* is somewhat higher and the maximum request range is by 5 cm lower than the expected 50 cm range, is attributed to about 2 dB loss in the RF detector and some mismatch loss at the Rx antenna.

With 508 µW of dc power, the maximum request range of the 2.49 GHz SO at 17 dBm of request power was reduced to about 20 cm which is also in a good agreement with (1).

The dependence of the SO supply voltage versus interrogation distance is shown in Fig. 6 for the 1 GHz SO. The supply voltage under full load, including modulation flip flop, varies by a factor of 2.8 when the distance to the request unit varies between 45 and 5 cm. Despite the moderately high Q of the STW resonator, a certain dependence of the SO frequency on the request power proportional supply voltage and, therefore, the distance to the interrogation unit, as well as on the propagation properties of the RF channel will be expected. This frequency variation is shown by the data plot in Fig. 7 which was measured in the overwrite mode of the spectrum analyzer by moving the interrogation unit over the entire SO operation range, at different excitation angles and directions and placing different RF reflecting objects inside and around the propagation path. It is evident that the worst case uncertainty of 48.8 KHz in the remote measurement of the SO frequency, does not exceed ±25 ppm which is by a factor of 10 less than the uncertainty of a directly interrogated SAW based SPR sensor operating at three times lower frequency [4]. This is a clear advantage of the proposed SO principle compared to traditional wireless sensors.

B. The 2.49 GHz NRO Based Sensor Oscillator

As explained above, the uncertainty in the remote measurement of the SO frequency, according to Fig. 7, is not caused by the measurand and represents, therefore, the actual error of the RF-powered sensor oscillator. Although this error is by at least an order of magnitude lower, compared to the error of directly interrogated SAW devices, it can be even further reduced by increasing the Q of the sensing device stabilizing the SO. This is evident from Fig. 8 which represents the maximum frequency variation of a 2.49 GHz SO, operated as a NRO and stabilized with a SPR device as the one from Fig. 5, featuring a Q of about 8000. To efficiently transmit the second overtone of the 1.247 GHz SO frequency in the standard 2.4 to 2.5 GHz range, in this experiment, the Tx antenna in Fig. 1 was matched to the 2.49 GHz frequency and the request power was increased to 27 dBm. The data in Fig. 8 indicates a total of 54.4 KHz SO

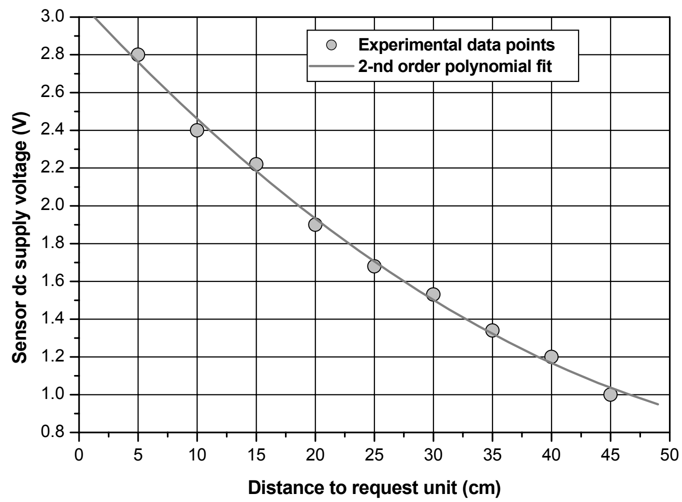


Fig. 6. Supply voltage versus distance to the request unit of the 1 GHz SO from Table 1.

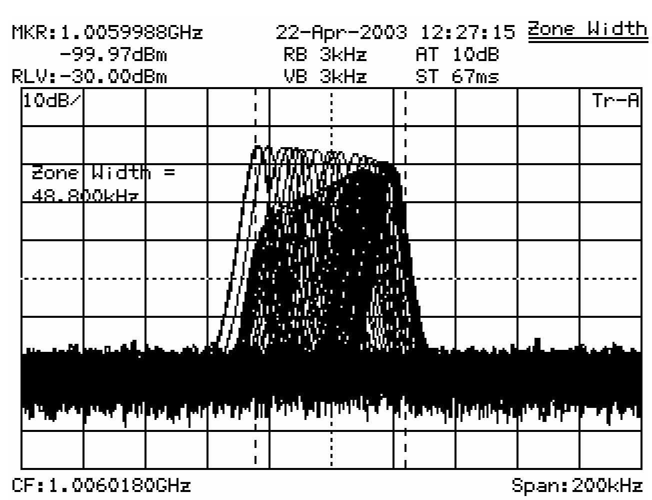


Fig. 7. Worst case frequency variation limits (dashed zone) of the 1 GHz FLO based SO over its operation range at 17 dBm request power. Q ≈ 3000.

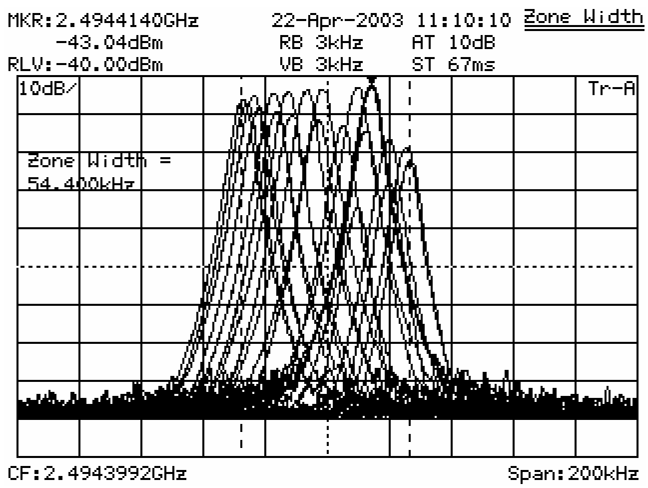
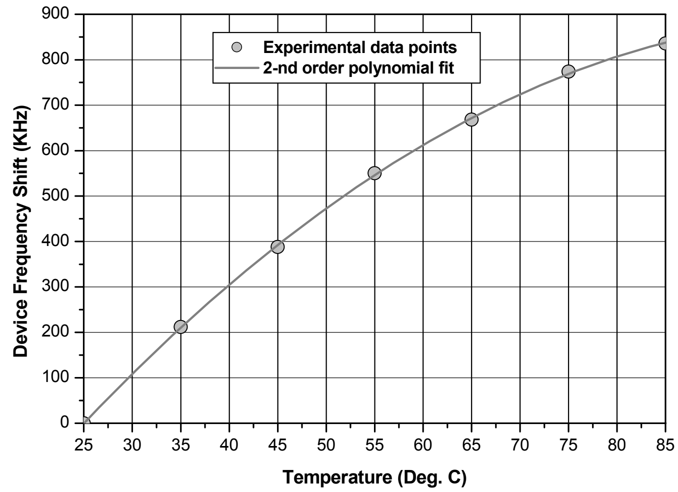


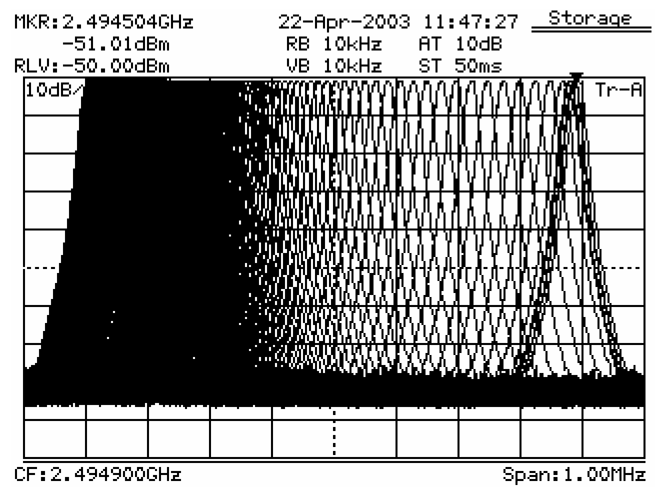
Fig. 8. Worst case frequency variation limits (dashed zone) of the 2.49 GHz NRO based SO over its operation range at 17 dBm request power. Q ≈ 8000.

frequency variation which results in an uncertainty of ± 11 ppm. This is an improvement by a factor of 2.3 compared to the 1 GHz SO with a Q of 3000 (see Fig. 7). In the author's opinion, even a further improvement is possible if the design of the sensor SPR is pushed close to the Q limit which is about 40,000 for STW based SPR devices operating at 1 GHz [12].

As shown in Fig. 9 a), the 1.247 GHz SPR device from Fig. 5, was fabricated on a quartz cut with a moderately strong temperature dependence around room temperature and, in this study, it was used as a sensor in a remote wireless temperature measurement. The data in Fig. 9 b) is the temperature induced frequency shift of the remotely interrogated 2.49 GHz NRO based SO, stabilized with this device. In this experiment, the SO was heated from room temperature to about 80 deg. C during interrogation with a 915 MHz CW power. The temperature induced SO frequency shift of 800 KHz from Fig. 9 b) is in an excellent agreement with the STW device temperature-frequency characteristic from Fig. 9 a) meaning that the SO exactly follows the temperature dependence of the



a)



b)

Fig. 9. Temperature behavior of a): the STW based SPR device from Fig. 5 and b): the 2.49 GHz NRO based SO using that device.

sensing STW device in the oscillator circuit. Taking into account the worst case frequency error from Fig. 8, the total uncertainty in this measurement is less than ± 2 deg. C. By using a temperature sensitive cut orientation of the STW device or simply moving the turn-over temperature far away from the temperature range of interest [9], and by further increasing the oscillator Q, the measurement uncertainty can readily be improved by at least an order of magnitude.

C. The Receiver Circuit

Traditional wireless sensors require sophisticated radar-like transceivers, capable of separating the strong request signal from the weak sensor response [1-3], and complex digital signal processing circuitry for extracting the measurand data. If the request and response signals are not well enough separated in time which, in most cases, requires fairly large SAW propagation delays in the sensing device [2], then the sensor response is masked by multiple-path signals and error probability increases dramatically. In the author's opinion, the proposed RF-powered SO concept offers an additional significant advantage compared to traditional wireless sensors and this advantage is in the simplicity of remote reception and sensor data processing with a greatly reduced error probability.

The block diagram of the simple portable 1 GHz double conversion superheterodyne receiver, used in this study, is shown in Fig. 10. The FSK modulated 1 GHz sensor response signal, generated by the SO from Fig. 1, is received by the input antenna, amplified by a low-noise preamplifier and downconverted by Mixer 1 to a 1-st IF in the 20 to 30 MHz range using a fixed-frequency temperature compensated 1027 MHz STW based 1-st local oscillator (LO) with a few ppm stability over the temperature range of (-20 to +70) deg. C. Thus, the frequency error, introduced by the 1-st LO, is about an order of magnitude lower than the SO uncertainty according to Fig. 7. In power-line operated receivers the 1-st LO can be ovenized to even further reduce its error. After low-pass filtering, the measurand proportional 1-st IF is available for direct measurement by an external counter or processing by an external personal computer. Inside the receiver, the 1-st IF is further downconverted to a standard 10.7 MHz 2-nd IF by

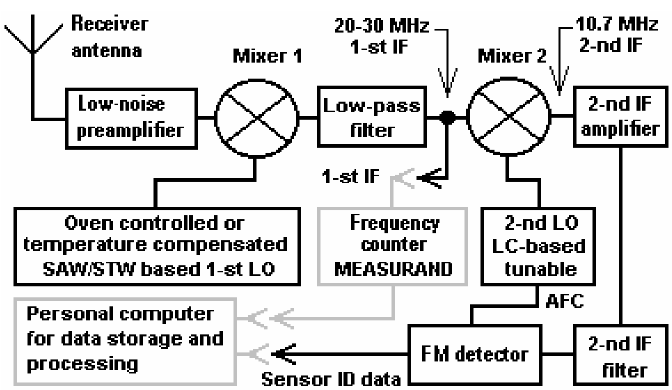


Fig. 10. Block diagram of the 1 GHz double conversion superheterodyne receiver for remote reception of the SO signal and retrieving modulation data.

Mixer 2 and a 2-nd LC-based tunable LO. After subsequent filtering and amplification, the 2-nd IF is FM detected and the sensor ID information, as well as possible additional low-frequency sensor data, are retrieved for processing by external counters or computers. In portable units, the 1-st IF and low-frequency data can be directly converted into measurand units and ID information and displayed on the front panel using low-power CMOS circuitry. Once the receiver is manually or automatically tuned to the SO frequency, which can be anywhere within in the 10 MHz reception bandwidth, the FM detector generates an automatic frequency control (AFC) voltage which makes the 2-nd LO track the SO frequency and keeps the receiver locked over a large measurand proportional variation range. This provides reliable FM detection even at a strong variation of the SO frequency during the measurement.

The following receiver performance was measured:

- Reception frequency range: 998 - 1008 MHz
- Reception bandwidth: 200 KHz
- Sensitivity for 12 dB S/N ratio 300 nV
- Supply voltage: 4.8 V
- Supply current: <40 mA

The receiver was tested with the 1 GHz FLO sensor from Table 2 which was interrogated at a 20 cm distance by 17 dBm of request power. The generated -25 dBm SO signal was reliably received over a distance of more than 100 m with line-in-sight conditions and through 4 walls in a concrete building.

V. Summary and Conclusions

This paper has demonstrated the feasibility of the RF-powered SAW/STW based sensor oscillator offering a better alternative to existing SAW based wireless sensor technology especially in the 2.4 to 2.5 GHz frequency range in which directly interrogated passive SAW sensors do not work well. The proposed SO is stabilized with low-loss, high-Q surface wave resonators and, and for stable operation, it needs very small amounts of d.c. power, typically a few hundred μW, which are readily obtained through RF detection of the request signal, regardless of its carrier frequency. Thus the SO can be interrogated in one frequency band in which the allowed transmission power is higher and the sensor response signal can be transmitted in another frequency band with a lower permitted power. Once the SO starts, due to the high Q of the sensing SAW resonator, the oscillator frequency is highly coherent and depends to a large extent on the measurand and not on interference with multiple-path or other spurious signals. Although the SO power may be fairly low, it can be received over large distance by means of a simple superheterodyne receiver, and line-in-sight conditions are not necessary. In addition, the sensor oscillator can be FM, AM, FSK or PM modulated. This allows complex digital signal processing at the sensor location, using low-power CMOS circuitry, and simultaneous transmission of ID information and data from several measurands without using unique SAW devices. Continuous and pulsed power operation is possible.

In this study, several SO design concepts, operating at 1.0 and 2.49 GHz and using FLO and NRO were successfully tested. The FLO based SO provide excellent wideband FM and FSK modulation capabilities and are appropriate for transmission of not only ID information but also data from several measurands, obtained by nonacoustic sensors at the SO location. Due to the much higher oscillator Q, NRO based SO are capable of providing substantially lower uncertainty in wireless measurements of sensor frequencies proportional to varying physical quantities. In all SO cases, the frequency uncertainty due to strong variations of request power and antenna impedance in the RF-propagation channel was found to be at least an order of magnitude lower compared to traditional directly interrogated SAW resonator based passive sensors.

It is the author's belief that the proposed RF-powered sensor oscillator may be very competitive in a variety of wireless sensor applications in the 2.4 to 2.5 GHz frequency range.

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