CONF-9306-205-1

THE LEVER OSCILLATOR FOR USE IN HIGH RESISTANCE RESONATOR APPLICATIONS *

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JUL 27 1993
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Abstract

The Lever oscillator has been specifically designed for use with quartz resonator sensors. The use of quartz resonators as sensors is of particular interest and depending on the sensing environment, e.g., liquid, the oscillator design is both critical and difficult due to the wide dynamic range of resonator resistance possible due to damping of the resonator. Standard oscillator designs do not work well as sensor oscillators. An oscillator design will be presented that allows both frequency and loss (R_m) of the resonator to be determined over a wide dynamic range of resonator loss. The Lever oscillator uses negative feedback in a differential amplifier configuration to actively and variably divide (or leverage) the resonator impedance such that the oscillator can maintain the phase and gain of the loop over a wide range of resonator resistance.

Introduction

Standard oscillator designs, like the Pierce or Colpitts type, are good for frequency source oscillator designs which provide low phase noise and high over-all frequency stability. These designs do not necessarily work well for sensor applications because of their inherent phase and gain sensitivity to resonator loss, R_m . Operating the resonator at f_s is a good initial approach if one is to design an oscillator for use with a high loss resonator. At f_s the resonator is approximately a real impedance and is at or near the minimum impedance of the resonator. Figure 1 is the model for a bulk wave resonator, with f_s defined as the resonance of L_m and C_m . Since a high loss resonator is not likely to be highly inductive at any frequency, it would be undesirable to try to make the sensor operate at an inductive impedance as this will limit the dynamic range of the oscillator. It is also desirable to

design this type of oscillator with one side of the resonator at a.c. ground. With one side of the resonator grounded the parasitic capacitances due to the fluid can be more easily controlled; this is especially true if only one side of the resonator is used for liquid exposure.

Figure 1. The quartz resonator model.

Design Goals

There are a few oscillator configurations which operate the resonator near f_s and have one side of the resonator grounded. None of these configurations meet the design goals of operating at f_s of the resonator and outputting a voltage proportional to the resonator loss R_m , by way of an automatic level control (ALC) circuit. For *in situ* resonator sensors the R_m of the resonator will have a wide dynamic range. This dynamic range depends upon the resonator and liquid properties but could be tens of ohms in air to several thousand ohms in a viscous liquid. The resonator loss is a function of the viscosity-density product (1). Since quartz resonator frequency shifts due to a given environment are

* This work performed at Sandia National Laboratories by the U.S. Department of Energy under contract DE-AC04-76DP00789

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determined by calculating the change in f_s of the resonator, it is convenient to design the oscillator to operate as close to f_s as possible such that the oscillator frequency mirrors the sensor-related frequency changes. The ability to operate at exactly f_s of a resonator is in itself a difficult task and the author has found that on an absolute basis it is impossible. In general if the impedance of C_0 is relatively large compared to R_m then f_s is close to the frequency at which the resonator impedance phase is zero. The R_m output voltage can also be used to determine the frequency error of the oscillator itself relative to true f_s if the shunt capacitance across the resonator is relatively constant throughout the measuring medium.

The Oscillator Design

Figure 2 is schematic of an oscillator that evolved into what is called the Lever oscillator. Understanding this circuit will help in understanding how the design goals were ultimately met.

Figure 2. The non-inverting amplifier as an oscillator.

In this oscillator the collector of Q_2 is directly connected to base of Q_1 and the resonator is connected from the base of Q_2 to ground. The resonator will control (poorly) the gain and phase of the potential oscillation. The current source I is controlled by V_c and will be used in an ALC circuit to servo a given oscillation amplitude and thus will reflect the resonator loss. Transistor h is the reciprocal of transconductance g_m and is equal to $(26 \text{ mV})/I_e$ where I_e is the individual transistor bias current. To determine the loop gain equations one could open the loop at the base of Q_1 and determine the loop gain equation for this amplifier, Figure 3. If the output impedance of the circuit is low relative to the input impedance the gain equation will be fairly representative of the actual oscillator. Equation (1) is the open loop gain equation for the circuit in Figure 3. This circuit, when V_{out} is connected to V_{in} , will oscillate when A_v is greater than or equal to one.

Figure 3. Open loop circuit for oscillator calculations.

If Q_1 and Q_2 are biased at reasonable current levels ($0.1 < I < 10 \text{ mA}$), the resonator will weakly control the oscillator because the resonator impedance is divided by the transistor β . Since the loop gain is more dependent on transistor h than the resonator impedance the oscillator sensitivity to resonator resistance is poor and the loop phase of the circuit will not be very sensitive to the resonator impedance. This circuit would work well for the application described if the resonator impedance was effectively divided by a smaller number (less than β) and was independent of transistor β . For example it would be desirable to have the resonator impedance divided by a large number when the resonator is a high impedance and divided by a smaller number when the resonator is a low impedance. This variable divider would help to compress the dynamic range of the oscillator transconductance over a wide dynamic range of resonator loss while still allowing enough sensitivity to determine the resonator loss via an automatic level control (ALC) circuit. This function can be obtained by modifying the oscillator of Figure 3 by placing a feedback resistor from the collector of Q_2 to the base of Q_1 . The Lever oscillator design (open loop) can be seen in Figure 4. This schematic shows the feedback resistor, R_f , that is used to make the oscillator sensitivity to resonator impedance a function of the resonator impedance itself. This function can be made β independent and will allow the resonator impedance to be effectively divided (leveraged) as a function of the magnitude of the resonator impedance.

Figure 4. The Lever oscillator open loop circuit.

The Lever oscillator open loop gain, A_v if $R_c < R_f$:

$$A_v \approx \frac{R_c}{2h + \frac{Z_{res} \parallel [R_f + R_c]}{\beta} + \frac{R_c Z_{res}}{R_c + R_f + Z_{res}}}; \quad (2)$$

for oscillation A_v must be greater than or equal to one. If all the variables in equation (2) are real, then Z_{res} must also be real (have a zero impedance phase). If Z_{res} is real then the frequency of operation is close to f_s of the resonator and the resonator impedance can be approximated by R_m . Equation (2) can be simplified to:

$$A_v \approx \frac{R_c}{2h + \frac{R_c R_m}{R_c + R_f + R_m}} \geq 1 \quad (3)$$

where β is considered large enough to make the center term in the denominator of the loop gain equation (2) very small in comparison to the other two terms. The term

$$\frac{R_c R_m}{R_c + R_f + R_m} \quad (4)$$

is the "Lever" term, since it will determine how the resonator impedance is effectively divided (leveraged) by the feedback circuit elements, R_c and R_f . R_m is effectively divided by the function

$$\frac{R_c + R_f + R_m}{R_c} \equiv g(R_m) \quad (5)$$

which is small when the resonator impedance is small and large when the resonator impedance is large. The Lever term (4) is transistor β -independent and can be made to dominate the given gain equation (2) over a wide range of β . This variable leverage function allows the oscillator to operate over a wider dynamic range of resonator resistance while

maintaining sensitivity at lower resonator resistances than if the divider effect was constant or dependent on β . Figure 5 shows a graph of the equation (5) vs R_m . This graph represents the variable divider effect the circuit has on R_m with R_c equal to 150 ohms and R_f equal to 1000 ohms. Due to the fact that the circuit is operating close to f_s of the resonator, Z_{res} is approximately R_m if C_0 is relatively small.

Figure 5. Graph of variable "Lever effect", $g(R_m)$ vs R_m .

For the values of R_c and R_f chosen, the resonator impedance R_m is effectively divided by 7 for small values of R_m and by approximately 24 for large values of R_m . By selecting the values of R_f and R_c such that $R_f/R_c \ll \beta$, the oscillator will be relatively independent of β . The choice of R_f and R_c in combination with the current source function (V_c) allow the designer to set the oscillator sensitivity to a desired range of R_m .

Lever Oscillator Practical Realization

The oscillator theory presented to this point has virtually ignored transistor capacitances and the practical matters of the design. Figure 6 is a schematic of a practical form of the Lever oscillator.

Figure 6. Lever oscillator schematic.

This schematic will be used to describe and derive the actual design performance of the oscillator function. In the theory stated above it was stated that if all the circuit components shown (Figure 4) were real then at the oscillation frequency the resonator would also be real. But in actuality, at frequencies greater than d.c., circuit capacitance will influence circuit performance. Capacitance from collector to ground of Q_2 combined with base to ground capacitance of Q_1 must be addressed. This capacitance, lumped together because of the circuit configuration, significantly degrades circuit performance: (1) this capacitance pulls the oscillator low in frequency (even for a small amount of capacitance) and (2) if the resonator becomes highly lossy, the circuit has a tendency to jump to a parasitic (capacitor controlled) frequency. These two problems were minimized by placing a parallel L-C (L_C and C_C) tank circuit across R_C . This tank is designed to be low Q (1 to 3) and is used to make the collector to ground impedance at Q_2 to be real (approximately R_C) at the desired frequency and to help constrain the oscillation frequency to the resonator and not to a parasitic frequency. This tank is adjusted such that with a resistor (low value, 100 ohms) in place of the resonator the tank circuit is adjusted to make the oscillator frequency approximately equal to the desired resonator frequency. With R_C real at the resonator frequency we have satisfied one of the assumptions given in the theory above.

The other assumption in the theory is that R_f is also real. Since R_f is a major component in the design, it is interesting to see the effect of capacitance across R_f due to the C_{bc} of the transistor. By using equation (3), substituting Z_{res} for R_M and Z_f for R_f one can solve for Z_{res} and determine the phase and magnitude of Z_{res} as a function of Z_f , R_C and h .

$$Z_{res} \approx \frac{[R_C - 2h][R_C + Z_f]}{2h} \quad (6)$$

and if $R_C < Z_f$

$$Z_{res} \approx \frac{[R_C - 2h][Z_f]}{2h} \quad (7)$$

where R_C and h are real. If these assumptions are good then it can be seen from equation (7) that the resonator impedance phase must be equal to the phase of Z_f . If operating close to zero impedance phase of the resonator is desired it is important to select R_f such that R_f is much less than the impedance of C_{bc} for the chosen transistor. If one was to design this oscillator with a transistor having a C_{bc} of approximately 3 pF and R_f of 1000 ohms then the impedance, Z_f for R_f in

parallel with 3 pF at 6 MHz, will equal 994 ohms at -6.45° . For this case the resonator impedance phase at the oscillation frequency will also be approximately 6° . Although the equations derived throughout this paper are for small-signal low-excess loop-gain (one) conditions, the oscillator essentially behaves as equation (6) predicts; this is true for two reasons: (1) the oscillator will be used in an ALC circuit which is used to control the excess gain to a relatively low value so that R_M can be determined and (2) the differential amplifier limiting characteristics provide a symmetrical limiting action that makes the circuit behave more linearly than a simple single-transistor oscillator circuit. Figure 7 is a calculation of resonator impedance and phase versus transistor h using equation (6).

Figure 7. Calculated resonator magnitude and phase vs h calculations for the Lever oscillator.

These graphs demonstrate the wide dynamic range of resonator impedance possible with the oscillator controlling the crystal impedance phase to approximately the phase of the feedback impedance Z_f .

Lever Oscillator Performance Data

To indirectly measure resonator resistance an ALC loop was designed to control the oscillator gain (h) via the voltage controlled current source, Q_3 , (Figure 8) of the oscillator.

Figure 8. The Lever oscillator with ALC circuit.

The ALC was designed to control the oscillation amplitude at the base of Q_1 to approximately 200 mV peak-to-peak. Figure 8 also shows an inductor L_x which can be used to tune out static capacitance of the fixture and resonator. L_x can dramatically increase the upper resistance limit before the oscillator breaks into a parasitic oscillation. To evaluate the Lever oscillator for one set of circuit parameters, a 6 MHz test resonator was made using a standard low resistance device in series with added resistance. Figure 9 shows the test resonator used to evaluate the Lever oscillator.

Figure 9. Test resonator with variable resistance.

C_{shunt} was included in the test resonator to more accurately simulate the C_0 of a standard resonator. This resonator was used to determine the actual resonator impedance phase at oscillation for a wide range of resistance, R_{test} . R_{test} was varied from 56 to 1700 ohms and the resonator impedance was measured on an impedance analyzer at the frequency of operation of the Lever oscillator for each tested resistance. Also the d.c. voltage of the ALC was recorded for all values of R_{test} . Data relating resonator impedance, phase, and d.c. output voltage as a function of R_{test} are presented in Figures 10 and 11 respectively.

Figure 10. Measured resonator impedance phase vs magnitude.

Oscillator V_{out} (d.c.) vs Resonator Resistance

Figure 11. ALC voltage out vs resonator resistance.

These data demonstrate that the oscillator does indeed servo the resonator impedance to approximately the phase of Z_T , in this case approximately 6° , when the resonator is above 200 ohms. At resonator impedances below 200 ohms the resonator impedance phase increases. This phase shift at low resonator impedances is explained by the fact that the oscillator is less controlled by the resonator when the resonator impedance is low than when large and is therefore more sensitive to circuit parameters not used in the approximations presented here. To illustrate this, a more exact model of the oscillator was used which assumed the transistor h to be very slightly capacitive. When this capacitance is included in the calculations the resonator impedance phase increases with decreasing resonator impedance just as seen in the data, Figure 10. Figure 12 shows the calculated magnitude of Z_{res} versus the phase of Z_{res} with transistor h slightly capacitive.

Figure 12. Calculation of resonator impedance phase vs magnitude for h slightly capacitive.

To demonstrate the β insensitivity of the Lever oscillator, a negative resistance model was used to calculate the input impedance (Z_{in}); the calculations are shown in Figure 13 as a function of h for transistor β 's of 150 and 50.

Figure 13. Lever oscillator circuit used in negative model calculations.

Figure 14 is a graph of the negative resistance function of the lever oscillator for β equal to 50 and 150. The oscillator β sensitivity is greatest at high resonator resistances where the negative feedback due to R_f and R_c is least effective. With the feedback used in this example the oscillator is relatively insensitive to transistor β for h between 20 and 65 ohms.

Figure 14. Calculation of β sensitivity vs h .

Oscillator Frequency Error

Now that it has been established that the Lever oscillator function can be made to act like an impedance phase servo it would be interesting to determine how this phase frequency relates to the true f_s of the resonator. If C_0 is zero then f_s is equal to the zero impedance phase frequency. But since C_0

is never zero then the oscillation frequency of the Lever oscillator will be offset from true f_s of the resonator by some amount depending on R_m , C_0 , and the phase of the resonator impedance which the oscillator is servoing. Figure 15 is a graph showing the calculated difference between the frequency of zero impedance phase of the resonator and f_s of the resonator. This is done for two static capacitances of the resonator, 1 pF and 3 pF. The graph shows that, for C_0 equal to 1 pF, the frequency error with C_m equal to 25 fF and L_m equal to 31 mH is relatively small but increases with resonator resistance R_m .

Figure 15. Frequency at 0° phase $-f_s$ vs R_m .

With increasing C_0 the frequency error increases dramatically with R_m . Figure 16 is a graph of the same function but in this case a difference frequency for -6° of impedance phase and f_s is calculated and the resulting curve shows a negative error for small R_m , sweeping positive as the resistance increases.

Figure 16. Frequency at -6° phase $-f_s$ vs R_m .

This is to be expected since at f_s the resonator is essentially a capacitance (C_0) in parallel with resistance (R_m). Therefore, the resonator impedance phase at f_s is a negative quantity varying with R_m and/or C_0 . To make a "true" f_s oscillator one would have to vary the phase that the oscillator serves as a function of R_m given a fixed C_0 . Due to the fact that the Lever oscillator has a R_m -sensitive output voltage, one can use this output to correct for the frequency error calculated in Figures 15 and 16 if such accuracy is required. This correction could be implemented using digital analog techniques.

Conclusions

A series resonant oscillator design has been presented that operates over a wide dynamic range of resonator resistance. This resistance range is compatible with *in situ* resonator sensors used in liquid environments. The Lever oscillator is useful for these applications because of its ability to measure the f_s and R_m of the resonator. Also, the oscillator's variable leveraging effect has been shown to be useful for increasing the dynamic range of resonator resistance while maintaining sensitivity at lower resistances. The oscillator's ability to servo a resonator-impedance phase has been demonstrated and this phase can be made close to zero degrees by selection of transistor type and/or circuit resistance values. The relationship between the resonator frequency at a given phase with respect to true f_s was discussed along with correction methods. Currently this design is being implemented in a jet-fuel contamination sensor system and in a viscosity-density sensor using bulk wave resonators (2). Variations on this design have worked with resonators with R_m in the 3500 ohm range, at approximately 6 MHz, and direct coupled, integratable, versions of this oscillator design are also being developed for applications where small size is important.

Acknowledgements

The author would like to thank Allyn R. Anderson for his assistance in the development of the Lever oscillator. His fabrication, testing and evaluation of the design helped to make the Lever oscillator a robust, successful design. Thanks also to Steve J. Martin and Elmer A. Klavetter for giving the author the design challenge that precipitated the development of the Lever oscillator and for their technical support and encouragement.

References

- (1) S. J. Martin, K. O. Wessendorf, C. T. Gebert, G. C. Frye, C. W. Cernosek, L. Casaus and M. A. Mitchell, "Measuring Liquid Properties With Smooth-And Textured-Surface Resonators," Proc. 47th Annual Frequency Control Symposium, June 1993.
- (2) T. J. O'Hern, W. M. Trott, S. J. Martin and E. A. Klavetter, "Advanced Diagnostics for *in situ* Measurement of Particle formation and Deposition in Thermally Stressed Jet Fuels," AIAA 930363, Aerospace Science Meeting, Reno, Nevada, 1993.

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