The Active-Bridge Oscillator*

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Introduction

This paper describes the Active-Bridge Oscillator (ABO), a new concept in high-stability oscillator design. The ABO is a bridge-type oscillator design that is easy to design and overcomes many of the operational and design difficulties associated with "standard" bridge oscillator designs. The ABO will oscillate with a very stable output amplitude over a wide range of operating conditions without the use of an automatic-level-control (ALC). A standard bridge oscillator design requires an ALC to maintain the desired amplitude of oscillation. For this and other reasons, bridge oscillators are not used in mainstream designs. Bridge oscillators are generally relegated to relatively low-volume, high-performance applications. The Colpitts and Pierce designs are the most popular oscillators but are typically less stable than a bridge-type oscillator.

The ABO is a highly integratable circuit and can be designed for use over a very wide frequency range with only minor circuit variations. The high Q of this design can dramatically improve phase-noise, reduce power-supply sensitivity, and minimize circuit related pulling problems common to most popular oscillator designs. The design intention of the ABO is *not* to make a state-of-the-art bridge-type oscillator but to provide a practical high-performance alternative that delivers most of the performance advantages of a bridge oscillator without the inherent disadvantages of complexity, design difficulty, high-power requirements, and integration potential.

Standard-Bridge Oscillator

Reference material for standard-bridge oscillators (SBO) can be found in papers written by Benjaminson[1,2] beginning in 1984. Reference 3 contains some historical information on bridge oscillator circuits. For the derivations that follow reference 1 will be used. The derivation of loop-phase is a general case solution where as reference 1 made a simple approximation yielding a specific case solution. Therefore different equations for loop-phase and ultimately loop Q result. Both the special and general case solutions will be described herein.

Figure 1 is a schematic of an SBO. The loop-gain equation for the oscillator is:

$$A = (\beta_p - \beta_n)A_v = 1 \angle 0° \qquad (1)$$

where A is the loop-gain. For oscillation A must equal 1.

Also, β_p must be greater than β_n for the phase condition of oscillation to be met. β_p is defined to be the positive feedback term and β_n is the negative feedback term.

For figure 1, $\beta_p = R_2 /(R_2 + R_1)$ \qquad (2)

and $\beta_n = Z_{res} /(Z_{res} + R_f)$ \qquad (3)

Z_{res} is the impedance of crystal resonator Y_1. The resonator impedance is be described by the following equation:

$$Z_{res} = R_m + jL_m\omega + \frac{1}{jC_m\omega} \qquad (4)$$

or Z_{res} can be written:

$$Z_{res} = R_m + j\frac{\omega^2 L_m C_m - 1}{\omega C_m} \qquad (5)$$

where R_m, C_m, and L_m are the motional parameters of the quartz resonator. This equation assumes that R_m is much smaller than the reactive impedance of the resonator shunt capacitance, C_0. This assumption is very good when using the resonator at series resonance. Since this is a series resonant oscillator, Z_{res} at series resonance is equal to R_m, the motional resistance of the resonator.
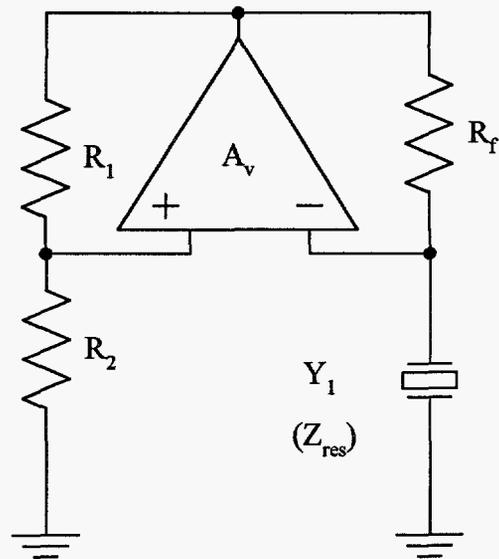
Figure 1. Standard bridge oscillator circuit.

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Hence the negative feedback term at series resonance is:

$$\beta_n = R_m / (R_m + R_f) \qquad (6)$$

If there is excess loop-gain and the loop-phase is zero then the nonlinear properties of the gain stage will limit A_v to a value commensurate with equation (1).

The term $(\beta_p - \beta_n)$ contains all the frequency-dependent loop-phase information. β_n has a phase versus frequency function dependent upon the values of R_f and Z_{res}, equation (3). This frequency dependent vector is subtracted from a fixed real quantity, β_p. The resulting vector, $(\beta_p - \beta_n)$, determines the oscillator loop-phase. As the magnitude of β_p approaches β_n, with $\beta_p > \beta_n$, $d\theta/d\omega$ of the resulting vector is greater than $d\theta/d\omega$ of β_n. The penalty paid for this increased phase-rate vector is that the magnitude of this resulting vector decreases, requiring larger A_v to sustain oscillation. Also, as the resulting vector becomes small it is more prone to 1/f noise in the sustaining amplifier.

The circuit loop-phase shift relative to frequency, ω, is adjustable by choosing the feedback ratios, β_p and β_n, such that the desired loop Q is achieved. To determine the effect β_p and β_n have on $d\theta/d\omega$ of the loop-phase an equation for the loop-phase versus ω is required. The loop-phase $d\theta/d\omega$ will be compared to the to the $d\theta/d\omega$ of resonator impedance (Z_{res}) phase to determine the relative loop Q. Let M_Q be the ratio of the derivatives of oscillator loop-phase to resonator impedance phase with respect to ω. M_Q is a measure of the loaded Q of the oscillator in that Q can be equated to the derivative of phase change relative to ω.

In reference 1 the $d\theta/d\omega$ for the resonator and the loop-gain are derived for figure 1. Resonator Q was correctly equated to $d\theta/d\omega$ in reference 1 by deriving the $d\theta/d\omega$ of the resonator impedance, equation (5). The following equation is the derivative of the resonator impedance phase with respect to omega:

$$\frac{d(\theta_{Resonator})}{d\omega} = \frac{2Q}{\omega_s} = \frac{2L_m}{R_m} \qquad (7)$$

where ω_s is the series resonance of the resonator.

$$\omega_s = \frac{1}{\sqrt{L_m C_m}} \qquad (8)$$

and $f_s = \omega_s / 2\pi$.

The loop-gain $d\theta/d\omega$ is found by taking the derivative of the phase of $(\beta_p - \beta_n)$ with respect to ω. The loop-phase equation is determined first by finding the phase of the following:

$$(\beta_p - \beta_n) = \beta_p - \frac{Z_{res}}{Z_{res} + R_f} \qquad (9)$$

Substituting the motional arm impedances for Z_{res}, $(\beta_p - \beta_n)$ becomes:

$$(\beta_p - \beta_n) = \beta_p - \frac{R_m + j\dfrac{\omega^2 L_m C_m - 1}{\omega C_m}}{R_m + j\dfrac{\omega^2 L_m C_m - 1}{\omega C_m} + R_f} \qquad (10)$$

Let θ_{Loop} be the phase of the expression $(\beta_p - \beta_n)$. After a little algebra, θ_{Loop} is:

$$\theta_{Loop} = \tan^{-1}\left(\frac{\dfrac{\omega^2 L_m C_m - 1}{\omega C_m}(\beta_p - 1)}{\beta_p(R_m + R_f) - R_m}\right) - \tan^{-1}\left(\frac{\dfrac{\omega^2 L_m C_m - 1}{\omega C_m}}{R_m + R_f}\right) \qquad (11)$$

The derivative of θ_{Loop} with respect to ω was found with a symbolic math program. The solution of $d(\theta_{Loop})/d\omega$ is lengthy and is simplified greatly by using the assumption that the oscillator operates at ω_s. At ω_s, $(\omega^2 L_m C_m - 1)/(\omega C_m) = 0$ and a relatively simple equation results. The following is the expression for $d(\theta_{Loop})/d\omega$ at ω_s:

$$\frac{d(\theta_{Loop})}{d\omega} = -2L_m\left(\frac{R_f}{(R_m + R_f)(\beta_p(R_m + R_f) - R_m)}\right) \qquad (12)$$

The goal of this exercise is to find the expression for M_Q which is the ratio of $d(\theta_{Loop})/d\omega$ to $d(\theta_{Resonator})/d\omega$. The following is the derived expression for M_Q:

$$M_Q = \frac{(\beta_n - 1)(\beta_n)}{\beta_p - \beta_n} \qquad (13)$$

Reference 1 derives a different but similar equation for MQ, equation (14). First, the sign of the two equations is opposite because reference 1 derived the phase of $1/(\beta_p - \beta_n)$. Upon careful examination of this derivation I also conclude that β_n was made proportional to R_m for reasons of simplification. If β_n is made proportional to R_m then this implies that $R_f \gg R_m$. This assumption simply forces $\beta_n \ll 1$.

$$M_Q = \frac{\beta_n}{\beta_p - \beta_n} \quad \text{(Benjaminson)} \qquad (14)$$

This M_Q is not the same as equation (13). When one assumes $\beta_n \ll 1$ then the magnitude of equation (13) does approach the magnitude equation (14), which would be consistent with my observations and conclusions. These facts explain and help verify the claim that equation (13) is a general case solution for the bridge

oscillator described in figure 1. Also, these results have been verified with SPICE analysis and negative resistance modeling. For any further reference to M_Q equation (13) is the assumed definition.

It is interesting to note that if β_p is equal to one, than $M_Q = \beta_n$. If $M_Q = \beta_n$ then the loop Q, $d\theta/d\omega$, can never be made higher than the Q of the resonator. This is because β_n can only approach one when R_f approaches zero. Although mathematically possible, this situation is impractical to say the least. By making β_p a little larger than β_n the circuit M_Q can be made ≥ 1. By proper selection of β_p and β_n M_Q may be designed for most any desired value around one with ease. Figure 2 graphically compares the loop-phase of a hypothetical 438 kHz oscillator with the resonator phase as a function of frequency. The feedback terms were: $\beta_n = 0.2$ and $\beta_p = 0.2857$. The resulting amplifier gain and M_Q were calculated to be: $A_v = 11.6$ and $M_Q = 1.87$. In this graph the phase of the resonator was inverted to better see the relative slopes of both curves, $d\theta/d\omega$. Note that at the oscillation frequency the slope of the oscillator loop-phase is greater than that of the resonator impedance phase by a factor of 1.87.
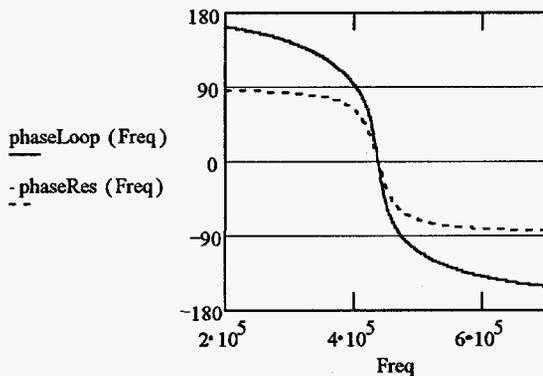
Figure 2. Oscillator loop-phase and resonator impedance phase versus frequency for $M_Q=1.87$.

Figure 3 is a plot of the magnitude of the loop-gain (A) versus frequency. For this example, the magnitude of A_v is 11.6 at the oscillation frequency of 438 kHz. This value of A_v corresponds to the minimum gain required for oscillation. This is the closed-loop steady-state value of A_v. The amplifier must have more open-loop gain than 11.6 to start-up in a reasonable amount of time and obtain a reasonable output amplitude. The gain, A_v will compress either by transistor nonlinearities, or by hard clipping to the value determined by the loop-gain equation (1), which is 11.6 for this example.

These two graphs demonstrate that the phase-slope of the oscillator loop is larger than the resonator phase slope and that the loop gain is at a local minimum at the oscillation frequency. Typically, non-bridge oscillators operate with the loop-gain at a local maximum.
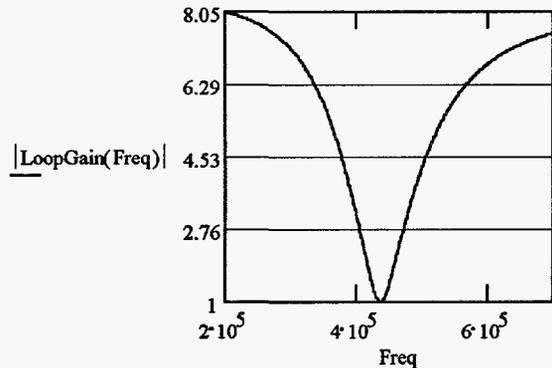
Figure 3. Oscillator loop gain versus frequency, $M_Q=1.87$.

Practical Realization of the SBO

Figure 4 is a practical realization of an SBO. This design uses Q_1 and Q_2 as the differential amplifier gain, A_v, where the collector of Q_1 drives the bridge-network. The bridge-network is comprised of C_1, C_2, R_f, and Y_1.
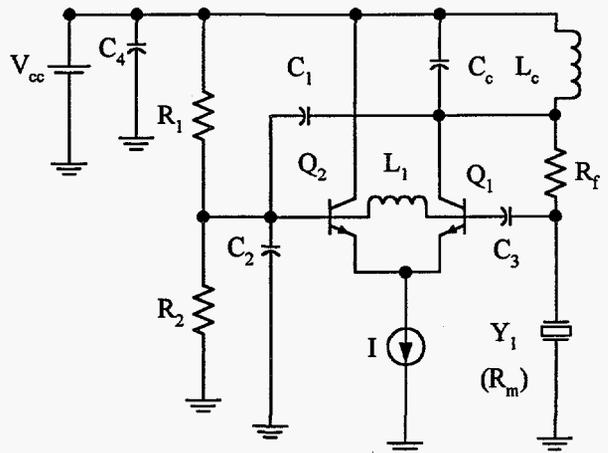
Figure 4. Schematic of a standard bridge oscillator.

L_1 is an RF choke that provides dc bias to the base of Q_1 equal to the base of Q_2. C_3 is a large value dc blocking capacitor. The tank in the collector of Q_1 (L_c, C_c) is used to make the collector impedance real at the desired resonator frequency. This SBO has some

inherent shortcomings and does not well represent the ideal bridge oscillator depicted in the "block" schematic, figure 1. First, this design uses a high-impedance collector node (Q_1) to drive the bridge network. This network uses C_1 and C_2 to provide β_p. R_f and Y_1 provide β_n. The differential amplifier gain is a function of the collector impedance and the transistor h where:

$$h = \frac{kT/q}{I_e} = 1/g_m \qquad (15)$$

$$\text{at } 25°C \qquad h = \frac{26mV}{I_e} = 1/g_m \qquad (16)$$

and I_e is the dc current in the transistor. The gain of the differential amplifier, A_v can be expressed as:

$$A_V = \frac{Z_{collector}}{2h} \qquad (17)$$

where $Z_{collector}$ is the impedance from collector to ground of the Q_1 transistor. Since R_f and the resonator directly load the collector the following equations can be used to describe the feedback and gain characteristics of the SBO in figure 4:

$$A_V \approx \frac{R_f + R_m}{2h} \qquad (18)$$

$$\beta_n = R_m /(R_m + R_f) \qquad (19)$$

$$\beta_p = \frac{\dfrac{1}{C_2}}{\dfrac{1}{C_2} + \dfrac{1}{C_1}} \qquad (20)$$

The A_v term is only an approximation and assumes no loading by the tank circuit (C_c, L_c, C_1 and C_2) or the bias resistors R_1 and R_2. Note that the tank circuit includes the β_p structure, C_1 and C_2. β_p is provided by the capacitive voltage divider between C_1 and C_2, equation (20). This equation assumes that R_1 and R_2 are much greater than the reactive impedances of C_1 and C_2.

To obtain relatively high M_Q it is desirable to make R_f relatively small, which provides significant negative feedback, β_n, equation (19). If R_f is made small A_v is decreased, from equation (18). The only way to overcome this loss in gain is to significantly increase the bias current in the differential amplifier transistors to decrease h, therefore increasing A_v. For ease of design and relative current efficiency it may be preferable to not have A_v a function of the feedback parameter impedances, in particular R_f. The only way this is achievable is to have a low impedance node at the collector Q_1.

Another disadvantage with the oscillator in figure 4 is that only the differential amplifier nonlinearities, h, control the amplitude of oscillation. As a transistor base-emitter ac voltage increases so does the effective nonlinear value of h. For high M_Q operation, the signals at each base are approximately the same amplitude. This means that the ac base-to-emitter voltages of the differential amplifier transistors are relatively small even for large signals at each base. Therefore, large changes in amplitude result in only small changes in operating h. This means that the oscillation amplitude will vary greatly with respect to resonator resistance, temperature, etc. Therefore this circuit must be used with an automatic level control (ALC) circuit which controls some element in the oscillator (bias or feedback component) continuously to maintain a desired signal amplitude.

The Active-Bridge Oscillator

The active-bridge oscillator (ABO) was discovered while modeling a half-bridge design. A half-bridge design is shown in figure 5.
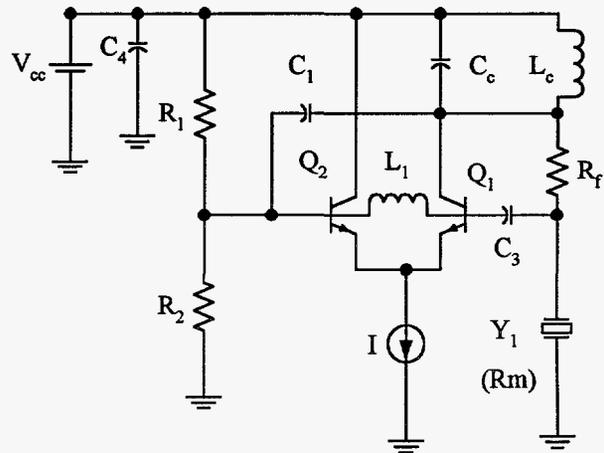
Figure 5. Schematic of a half-bridge oscillator design.

In the half-bridge design[4], β_p is set to one (C_1 is chosen to be large, ac short) and the oscillator uses only β_n to control the loop operating Q. Although not extremely stable, the design is very practical and has good amplitude stability. While modeling very high-frequency versions of this design the current source, I, was made non-ideal. In checking this model, relatively low-values of ac impedances were substituted for the high-impedance source. This modeling suggested improved stability with low ac impedances placed from the emitters of Q_1 and Q_2 to ground. Subsequent analysis showed a bridge effect in that as this emitter impedance

decreased the half-bridge design turns into a full-bridge function where β_p is now less than one. β_p becomes a function of this emitter to ground impedance and the transistor h in this non-obvious bridge circuit.

Figure 6 is the simplified active-bridge oscillator schematic used as a derivation reference.
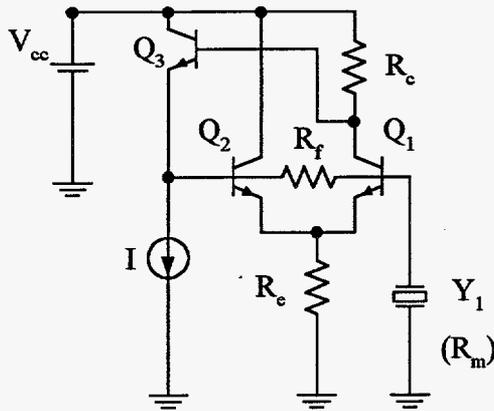
Figure 6. A simplified ABO schematic.

The loop equations can be derived by breaking the oscillator loop at the emitter of Q_3, base of Q_2 node and determining the loop-gain. This is a logical place to break the node and preserve the oscillator closed-loop characteristics. This is because the emitter of Q_3 is low impedance, and the β_p and β_n functions remain intact.

This circuit is designed with Q_3 used as an emitter follower to buffer the collector Q_1 from the low impedance β_n, and serve as an output buffer. β_n comprises R_f and Y_1. Y_1 is equal to R_m at ω_s. R_f also supplies dc bias to the bases of Q_1 and Q_2. This is possible because R_f is typically a relatively low value resistor, at most hundreds of ohms. R_e is low value resistor, tens of ohms (ac). R_e used in conjunction with the intrinsic emitter impedances of the transistors, h, to provide the β_p function. Hence the name active-bridge oscillator.

In deriving the loop gain, it was found to fit the form of a bridge oscillator function, equation (1), with the following characteristics. For simplicity these equations assume relatively large transistor β.

$$A_v = \frac{R_c}{h + \dfrac{hR_e}{R_e + h}} \tag{21}$$

$$\beta_n = R_m / (R_m + R_f) \tag{22}$$

$$\beta_p = R_e / (R_e + h) \tag{23}$$

β_p and A_v are functions of h and R_e. It is interesting to note that as R_e is decreased, increasing M_Q, that A_v is *increased*. Since increasing M_Q requires a larger A_v to sustain oscillation this characteristic is desirable from a current efficiency standpoint. Essentially, you get a little 'free gain' when it is required.

Figure 7 is a practical example of the ABO with consideration given to bias and filtering necessities.
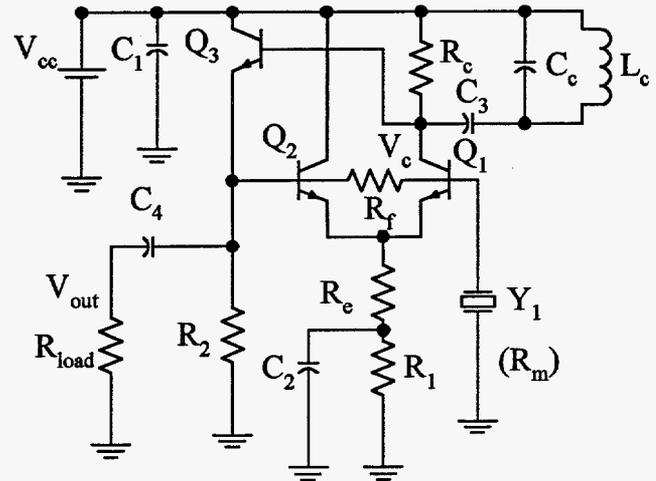
Figure 7. The complete active-bridge oscillator design.

This circuit includes R_1 and C_2 to facilitate dc biasing and provide an ac ground for R_e. Capacitors, C_3 and C_4, are dc blocking capacitors and considered as ac shorts. R_e provides β_p and is therefore a low value (typically $< 100\Omega$) where R_1 is a bias resistor and typically set from 1 kΩ to 2 kΩ. The bias voltages in the circuit are therefore primarily determined by V_{cc}, transistor V_{be}, R_1 and R_c. Equation (24) is the dc bias equation for the dc voltage V_c (figure 7) relative to ground. This equation is valid only when the circuit is not oscillating. When the circuit oscillates there is significant bias shifting in the circuit. V_c is the dc voltage at the collector of Q_1 when R_e is much less than R_1.

$$V_c = \frac{V_{cc} + 2V_{be}\dfrac{R_c}{2R_1}}{1 + \dfrac{R_c}{2R_1}} \tag{24}$$

This equation can be greatly simplified if $R_c = R_1$, which is a convenient and practical design possibility for this circuit.

$$V_c = \frac{V_{cc} + V_{be}}{1.5} \qquad (25)$$

when R_c is set equal to R_1, and $R_e \ll R_1$.

The emitters of the differential amplifier (Q_1, Q_2) will be biased at two V_{be} drops down from V_c due to the base-emitter drop of Q_3 and Q_2. R_2 sets the bias current in Q_3. The dc voltage at V_{out} is simply V_c minus V_{be}. Therefore the bias current for Q_3 is $V_{out}dc/R_2$. This bias scheme works nicely in that it is very simple yet biases all transistors simultaneously at reasonable voltages and currents. The tank circuit, L_c-C_c performs two functions. First, for efficient use of current and a high available A_v the collector resistor R_c, is chosen to be relatively large. Any significant circuit capacitance at this node will make the phase of the collector impedance a negative phase angle. For proper circuit operation all feedback terms and A_v need to be real. The tank, L_c-C_c, allows for A_v to be made real at any desired frequency. Secondly, this tank significantly reduces the distortion in the output waveform by shunting the undesired harmonics of the oscillator to ground. Also, this tank naturally selects the desired resonator overtone. Overtone operation at f_s is as easily achieved as fundamental mode operation.

Due to the predominately resistor-transistor based construction of this design, an analog ASIC would allow for a very small oscillator circuit that provides ease of design and high-performance over a wide range of operating frequencies. Reasonable values for an ABO when R_m=20 Ω are: R_c=R_1=1.5k Ω, R_e=75Ω, R_f=100 Ω and R_2=400 Ω. L_c and C_c are chosen to be parallel resonant and approximately 100 Ω to 150 Ω of reactive impedance. This results in an oscillator with high operating Q, low-distortion and a low-output drive at Q_3. This oscillator designed for 20 MHz operation delivers approximately minus 2dBm into a 50 Ω load. The waveform is a very low distortion (2nd harmonic –30 dBc) sine wave. This circuit required approximately 10 mA of current when designed for a V_{cc} of 5 V. Q_3 is the major power consuming component in that it must drive the low impedance R_f and R_{load}. To do this the bias in Q_3 is relatively high.

Active-Bridge Nonlinearities

The ABO steady-state, large-signal oscillation amplitude can be made relatively independent of resonator resistance, V_{cc}, and temperature changes. This is due primary to two factors. First, the steady-state base signals on the ABO are *not* very similar in amplitude as they are in a SBO. The base of Q_2 is the same amplitude as the collector signal at Q_1. The base of Q_1 is being driven by a fraction of the collector of Q_1. This allows the individual transistor h to be a strong function of signal amplitude. As the oscillation amplitude increases h increases. Secondly, the β_p term of the ABO is a strong function of oscillation amplitude, β_p= $R_e/(R_e+h)$ where h is the nonlinear transistor-impedance term. As a result the oscillation amplitude is relatively stabilized by the circuit M_Q via transistor h. Therefore slight changes in the circuit parameters, such as resonator resistance, do not effect the oscillation amplitude greatly.

These characteristics are useful in that the ABO does not require an ALC for low distortion large-signal sinusoidal outputs.

Figure 9 is a graph of a calculated M_Q versus transistor h for the following hypothetical ABO design: R_c=1100Ω, R_f=105Ω, R_m=20Ω, R_e=75Ω. Thus β_n calculates to 0.160. Using equation (1) and the oscillator feedback and A_v terms, equations (21), (22), and (23) the steady-state value for h was determined. The calculated value of h is 143 Ω. Therefore A_v= 5.6, β_p= 0.338 and M_Q calculates to 0.753.
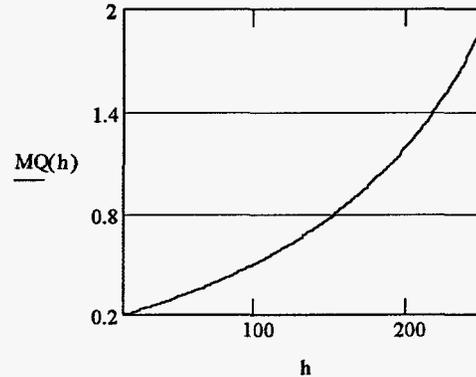
Figure 9. M_Q versus h for h equals 15 to 250 ohms.

Figure 10 is a graph of A_v versus h for the given hypothetical ABO. This graph demonstrates that A_v is decreasing with increasing h. It can also be seen in this graph that A_v is 5.6 at h = 143 Ω. This value of A_v is the steady state gain. A_v is equal to $1/(\beta_p-\beta_n)$, as equation (1) states.
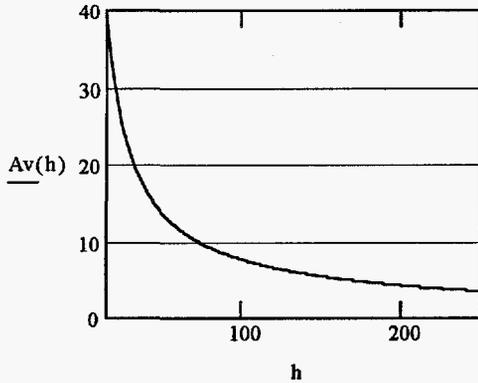
Figure 10. A_v versus h, for h equals 15 to 250 ohms.

To demonstrate the interdependence of A_v and the feedback terms this hypothetical oscillator circuit will remain constant and the resonator resistance will be allowed to vary slightly. Let the resonator resistance, R_m increases from 20 Ω from 25 Ω. The oscillation amplitude will decrease just enough to obtain a new steady-state value of h that satisfies the loop-gain equation (1). In this case h = 134 Ω and the resulting M_Q, feedback terms, and A_v are: M_Q = 0.936 β_p=0.358, β_n=0.192 and A_v=6.02. Note that β_p changes from 0.338 to 0.358 for R_m going from 20 Ω to 25 Ω. M_Q changes by 24.3% for this 5 Ω change in R_m. This relatively small change in M_Q is due to β_p increasing with β_n via the change in transistor h.

It is interesting to compare a hypothetical SBO (like Figure 4) to the ABO with the same feedback parameters allowing the resonator R_m to vary from 20 Ω to 25 Ω. Let β_p=0.358, and R_f=105 Ω and R_c=1100 Ω but assume that the collector impedance (Q_1) is buffered by a transistor (as in the ABO) such that the A_v for the hypothetical SBO is $R_c/2h$. This allows a fair comparison of the two designs in that both circuits now buffer the bridge-network from the high impedance collector node.

When this sensitivity study is evaluated one finds that for an R_m change of 20 Ω to 25 Ω that M_Q goes from 0.7534 to 1.0632, a 41% change. Transistor h goes from 98.12 Ω to 80.35 Ω. The percentage change in h required by the SBO to absorb this change in R_m is approximately six times greater than for the ABO. M_Q for the ABO is more stable over this change in R_m than an SBO. Also, since the "amplitude versus h" function is much steeper with the ABO than an SBO (as previously described) the ABO amplitude stability is much greater than that of an SBO.

When these two oscillators were built and evaluated for R_m varying from 20 Ω to 25 Ω, the SBO oscillation-amplitude changed 33%, and the ABO amplitude changed only 5 %.

ABO with Class-AB Follower Stage

If the ABO is designed for a large value R_{load} and or a R_f impedance, the dc current required by the circuit to perform the desired function can be relatively low. To increase the oscillator power-efficiency for low-values of R_f and R_{load}, the class-A, non-efficient emitter follower, Q_3, needs to be replaced with a more current efficient class-AB follower. Figure 11 is a schematic for the ABO with a class-AB stage as the follower circuit. This class-AB stage significantly decreases the dc bias required to swing large signals through low-impedance loads. This is because of the push-pull action provided by the PNP-NPN output stage. The class-AB stage consists of Q_3, Q_4, Q_5, Q_6 and a bias resistor R_2. The bias established by R_2 is mirrored in Q_4 and Q_3. This circuit can greatly increase the overall power efficiency of the ABO. The earlier design example required approximately 10 mA of supply current for approximately –2 dBm output power. The same design implemented with this approach, at 20 MHz, required 5 mA for the same output power into 50Ω, R_{load}. For this circuit R_2 = 2.6 k Ω and the idle current in the output transistors is 1 mA. This circuit is also highly integratable and can operate to high-frequencies.
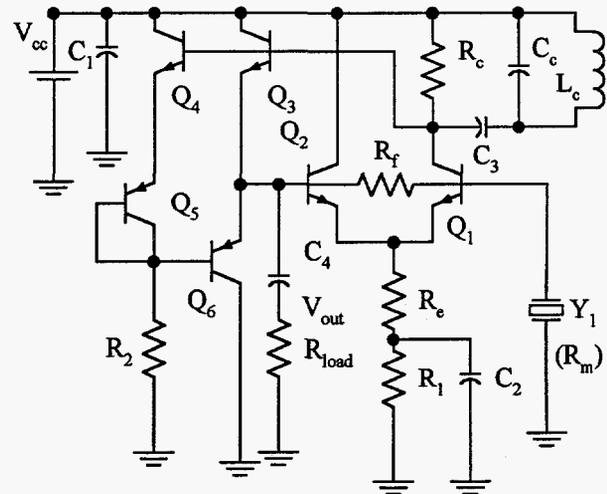
Figure 11. The ABO with a class-AB follower stage.

Active-Bridge Experimental Data

The ABO of Figure 11 was used to verify the ABO loop equations for M_Q. For this experiment R_c=1100 Ω, R_1=1500 Ω, R_f=105Ω, R_m=20 Ω, R_{load}=200 Ω. R_c was actually 1500 Ω but due to tank-loading and the finite impedance of the buffer, R_c=1100 Ω was used for the

calculations. A 20 MHz resonator was used for this test. The circuit M_Q was varied by changing R_e from 35 Ω to 75 Ω. The M_Q was determined by measuring the frequency shift of the oscillator when a small capacitor, 1pF, shunted the collector load of Q_1. This added capacitance places a small, known, phase-shift into the loop. Measuring the frequency shift required to accommodate the collector phase-shift measures the resonator phase-shift. These two phase-shifts allow for a determination of M_Q. The collector phase-shift for a 1 pF delta capacitance across R_c is approximately 7°. The resonator phase-slope measured 114.3 m°/Hz, near series resonance.

$$\text{Therefore: } M_Q\text{measured} = \frac{\left(\dfrac{7°}{\Delta\text{FreqOsc}}\right)}{\left(\dfrac{114.3\text{m}°}{\text{Hz}}\right)} \qquad (26)$$

where ΔFreqOsc is the frequency shift of the oscillator due to a 1pF capacitance placed in parallel with the collector tank of Q_1. Table 1 displays the results of the described experiment.

R_e	M_Q measured	M_Q calculated
35 Ω	1.33	1.23
51 Ω	1.13	0.957
75 Ω	0.87	0.75

Table 1. Data table for the M_Q experiment of an ABO.

These data demonstrate that the loop-phase indeed can be made steeper than the resonator slope, when R_e equals 35Ω. This data also demonstrates how well the large signal operation of the design can be predicted from the derived equations which are large signal approximations using small signal parameters, h. The oscillation amplitude decreased approximately 40% for R_e going from 75 Ω to 35 Ω. This is very good

performance considering the M_Q varies approximately 53% for this range of R_e.

The ABO With Fixed Feedback Terms

There are applications for bridge-type oscillators where it is undesirable to have $d\theta/d\omega$ a function of transistor h. Figure 12 shows a schematic of an ABO that is being used for a quartz-sensor application where the ABO function has a fixed value for β_p.

The impedance of Q_7 is substituted for the R_e term in the ABO loop-gain terms, A_v and β_p, of equation (21) and (23). Transistor Q_7 is configured as a diode. This diode impedance value is equal to the intrinsic emitter impedance. Q_1 and Q_2 are defined here to have intrinsic emitter impedance equal to h. Q_7 is biased at of twice that of Q_1 (or Q_2) making the impedance of Q_7 equal to h/2. Therefore, this oscillator has the following loop-gain terms where R_c is assumed to dominate the collector to ground impedance:

$$A_v = \frac{R_c}{1.5h} \qquad (27)$$

$$\beta_n = R_m/(R_m + R_f) \qquad (28)$$

Since $\beta_p = (h/2)/((h/2)+h)$

$$\beta_p = 1/3 \qquad (29)$$

β_p is now a constant and β_n is as before, a voltage divider between R_f and R_m. This circuit biases the collector of Q_1 to 3 V_{be} drops above ground. Q_6 sets the current in the differential pair, Q_1-Q_2 by current mirroring. The bias current in Q_5, controlled by Q_6 and R_1 is mirrored in Q_4. Due to the low value R_f, the bias current of Q_1 and Q_2 are equal to the current in Q_4.
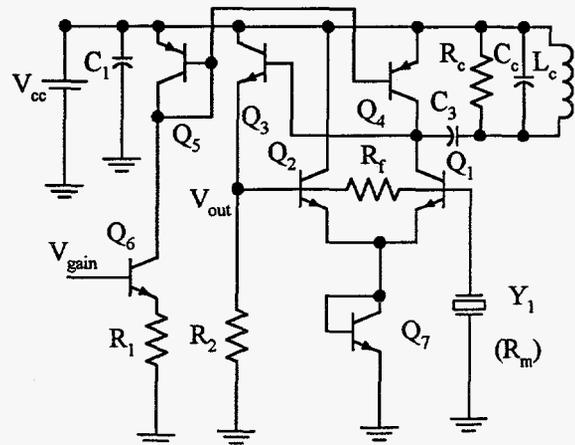
Figure 12. The ABO with fixed value feedback terms.

This current directly controls the magnitude of A_v via h, equation (27). Thus an ALC circuit can control the amplitude of oscillation of node V_{out} via a control voltage, V_{gain}. This design is currently being used to measure very small changes in resonator resistance where a quartz resonator is being used as a sensing element. In this application both frequency and R_m information are required. It is desired to resolve very small changes in R_m. This circuit has resolved milli-ohm changes in a nominal 25 Ω resonator. The original ABO circuit tends to mask these changes because β_p, is not a constant, as described earlier in this paper. Also to maintain circuit linearity, this design is being used at very small oscillation amplitudes. In this situation, the design behaves very similar to an ideal SBO. The main advantage of this design over an SBO is the circuit integration potential and the simplicity of the circuitry.

Conclusions

General bridge-oscillator design equations were derived and discussed as an introduction to a new oscillator, the Active-Bridge oscillator. The ABO feedback and gain terms were presented and used in hypothetical and actual design examples. These examples included comparisons to previous work, references 1 and 2.

This design is both practical, and in many ways an excellent general use oscillator that is good for high-stability applications at most any frequency that quartz resonators are constructed. ABO designs have been implemented easily with 125 MHz fifth overtone resonators or with fundamentals from 5 MHz to 200 MHz with essentially the same design! With high f_t transistors or ASIC implementations, higher frequencies could easily be achieved.

Another significant advantage of the ABO is that it operates with the resonator at series resonance. This is a distinct advantage at high-frequencies where manufacturing resonators tuned to f_s is easier and more consistent than when tuned to a given load capacitance. This is true because the ill effects of the resonator C_0 (shunt capacitance) are minimized with f_s tuning. At f_s the resonator is near a local low-impedance point which is relatively easy to measure and not very dependent on the value of C_0.

The ABO also produces a low-distortion sine wave output into low impedances with no additional filtering or buffering. Very efficient low-current designs are possible with the ABO using the class-AB follower. With V_{cc} as low as 3 Volts designs drawing 2mA at 20 MHz were easily achieved. The bias arrangement and nonlinear properties of the ABO allow operation over a wide temperature range with little or no amplitude variations. The design uses a grounded resonator which greatly simplifies the design of VCOs (voltage controlled oscillator) or TCXOs (temperature compensated crystal oscillator) that require some form of "pulling" circuitry such as varactor networks. The phase-noise of the design has been found to be at least as good as a well designed Piece oscillator using the same resonator. This design is still in development but it *appears* that the close-in noise is excellent and the far out (>10 kHz) noise is about 10 dB worse than a typical single transistor Pierce or Colpitts design. This is most probably due to the relatively large number of active components used to implement the oscillator. But, oscillator pulling due to temperature, power supply, or load, are superior to any oscillator (without an ALC) the author has yet encountered. The overall performance and design flexibility make this design a relative breakthrough in that it is nearly a "universal" design and can be used for many applications and frequencies.

References

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