VI-5 YIG TUNED AND VARACTOR TUNED L-BAND TRANSISTOR OSCILLATOR

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Voltage tunable oscillators covering the full octave range 1 to 2 GHz have been constructed using either a YIG resonator or a varactor diode as the variable element. It is the purpose of this paper to present a guide to the design of these oscillators and to compare their performance and their limitations.

GENERAL OSCILLATOR CIRCUIT

The general form of the equivalent circuit of a microwave transistor oscillator without the bias circuitry is shown in figure 1. Circuits of this form have been analyzed in the literature.[1,2] There are really two basic feedback mechanisms involved in this circuit, and the oscillator can operate with either one, or both, present. In the absence of any input capacitance, C_i, feedback is provided by collector-emitter capacitance C_{ce}, which is present in the transistor itself. On the other hand, if C_{ce} were absent, the capacitance C_i would be necessary for oscillation and, in this case, the feedback is through the base lead impedance which is common to the input and output loops. For widest tuning range in the frequency band near or beyond the f_T of the transistor, both feedback mechnisms have to be present. Tuning is accomplished by varying either X_i or X_L. In either case, optimum loading is achieved in the collector circuit. The loading network for broadband performance is, necessarily, more complex than the equivalent series impedance, Z_L, indicated in figure 1.

Although the complete circuit of figure 1 is readily analyzable by solving the characteristic equation, such a solution becomes rather involved, algebraically. For the present purpose, one can get a better insight into the performance limitations and the tuner requirements by examining approximate expressions for Z_i or Z_o. The impedance

$$Z_i \cong \frac{-k_i \omega L_b + j\omega L_b (k_r - \omega C_c X_L)}{1 - \omega C_c (\omega L_b + X_L)} + j\omega L_e \quad --- \quad (1)$$

The approximations are that r_b, R_L and C_{ce} are negligibly small. In the above expression, k_r and k_i are the real and imaginary parts of $(1 - \alpha)$. The values are given very approximately by

$$k_r = \frac{\frac{f^2}{f_T^2}}{1 + \frac{f^2}{f_T^2}} \quad \text{and} \quad k_i = \frac{\frac{f}{f_T}}{1 + \frac{f^2}{f_T^2}} \quad -- \quad (2)$$

Typically, f_T is of the order of 1.5 GHz for transistors which are suitable for L-band oscillators. The expression for Z_i will have a net negative real part and positive imaginary part over the frequency range, provided that the sum of the base and load reactances is small enough to be below resonance with the collector capacitance. The impedance

$$Z_o \cong r_b + j\omega L_b - \frac{k_i}{\omega C_c} \frac{C_{ce}}{C_c} - \frac{j}{\omega C_c} \quad --- \quad (3)$$

when $C_i = 0$ and C_{ce} is small compared to C_c. Z_o will have a net negative real and imaginary part over a wide frequency range. Note that, in this case, the existence of the negative real part is not affected by the resonance condition between base inductance and collector capacitance when C_i is zero.

YIG TUNED OSCILLATOR

A YIG sphere, coupled to a wire loop, behaves as a parallel resonant circuit with a constant external Q dependent upon the degree of coupling. The self-inductance of the loop, which can be relatively large, appears in series with the resonant circuit. This means that the tuner can positively control the frequency only if it is used in the oscillator circuit at a point where capacitive reactance is required for oscillation. In the circuit of figure 1 it is best used in the place of C_i, with one end of the loop connected to the emitter and the other end grounded.

Q measurements were made on a number of YIG spheres and coupling structures using a two-port test fixture in a 50 ohm system. Q_u and Q_e were calculated from measurements of resonance insertion loss and bandwidth at different resonant frequencies set by the dc magnetic field. Figure 2 shows the data for the structure used in the 1 to 2 GHz oscillator. This was considered optimum on the basis of best Q_u/Q_e ratio and moderate self-inductance. In operation in the oscillator, the YIG resonator is pulled off resonance sufficiently far to more than tune out the self-inductance and provide capacitive reactance at the transistor. The tuner performance is best specified then, by its effective series resonant Q. This is, approximately, $\frac{50}{\omega L_s}\frac{Q_u}{Q_e}$
where L_s is the series inductance of the coupling loop. For the configuration of figure 2 with $L_s \approx 3$ nanohenries, the effective tuner Q over the band is of the order of 15. This is important since it must be much greater than the negative Q of the transistor input impedance.

Since transistors with relatively high dissipation ratings have relatively large lead inductances as well as large collector capacitances, a low power transistor was used in the YIG tuned oscillator – Texas Instruments type SS7285. This transistor has an f_T of approximately 1.7 GHz, a base resistance of the order of 3 ohms, a collector capacitance of the order of 1 pf, lead inductances of the order of 1 nanohenry, and a dissipation rating of the order of 250 mw. The package is an epoxy-sealed pellet-type with nonmagnetic leads.

The design of the broadband leading network and the optimizing of base lead inductance are based upon measurements made with the YIG tuner coupled to the emitter and a tunable load at the collector. Contours of constant power at constant frequency, with variable load impedance, are plotted on a Smith Chart for several frequencies in the band and several base lead lengths. A load network is designed with an impedance function to intersect appropriate contours over the octave frequency range. The final oscillator circuit is shown in figure 3 and the power output vs frequency curve for the oscillator is shown in figure 4. Tuning linearity is of the order of a few tenths of a percent. The efficiency of the broadband load coupling network averages about 30% based upon the following measurements of maximum power output with the tunable load -- 90 mw at 1 GHz, 60 mw at 1.5 GHz, 30 mw at 2 GHz.

VARACTOR TUNED OSCILLATOR

Varactor diodes which have been used in L-band oscillators typically have Q's of the order of 25 at 1 GHz and –1 volt bias, while, at 2 GHz and –60 volts bias, (near reverse breakdown) Q's will approach 100. From zero external bias to 60 volts, the capacitance variation is about 7 to 1. Diodes with higher breakdown voltage, and resultant high tuning ratios, have substantially lower Q's and have not proved successful. Varactor tuning has been used with high dissiptation microwave transistors with the goal of producing relatively high power output of the order of 100 mw over the octave range. With such transistors the optimum position for the tuner has been at the collector, or the X_L position in figure 1. The tuner consists of a varactor diode in series with a fixed inductor. A parallel resonant tuner at this point would not allow as wide a tuning range as the series circuit tuner.

In the approximate expression of equation (3), the ωL_b term can be considered part of the fixed tuner inductance, so that the tuner must present an inductive reactance given by $\omega L_T - \frac{1}{\omega C_T} = \frac{1}{\omega C_e}$
where L_T is the total fixed tuner inductance (including L_b) and C_v is the capacitance of the varactor. One can derive the relation

$$\frac{C_{v_1}}{C_{v_2}} - \frac{\omega_2^2}{\omega_1^2} = \frac{C_{v_1}}{C_c}\left(\frac{\omega_2^2}{\omega_1^2} - 1\right) \qquad --- \quad (4)$$

where subscript 1 refers to the low frequency end, and subscript 2 to the high frequency end of the tuning range. For a 7:1 varactor capacitance ratio and an octave frequency range,

$$\frac{C_{v_1}}{C_c} = 1 \quad \text{and} \quad \omega_1 L_T = \frac{2}{\omega_1 C_c}$$

These conditions are quite relizable with high dissipation transistors having collector capacitances of the order of 5 pf. The problem exists with tuner losses at the high frequency end of the band. Here the real part of Z_o is

$$\frac{1}{\omega_2 C_c}\left(-k_i \frac{C_{ce}}{C_c} + \omega_2 R_b C_c\right)$$

For available transistors, $r_b C_c$ is of the order of 3.5 pico sec, k_i (at 2 GHz) is of the order of 0.6 when f_T is 1.5 GHz, C_{ce}/C_c is of the order of 0.2. This gives a negative output resistance of the order of $0.08/\omega_2 C_c$. The series resistance of the tuner is $1/\omega_2 C_{v2} Q_{v2}$ which is of the order of $0.07/\omega_2 C_c$ (with $Q_{v2} = 100$). The net negative resistance is not sufficient to allow any loading of the oscillator at this frequency.

The value of C_{ce} could be increased by adding capacitance externally, but this has not been very successful. However, addition of appropriate capacitance C_i at the emitter increases the high end negative resistance considerably without affecting the reactive tuning conditions very much and without affecting the low end negative resistance. This has been proved theoretically by analyzing the appropriate Z_o expression, and is born out experimentally. The transistor package must have a very small emitter lead inductance, and base lead inductance. The adjustment of the latter (external to the package) is a critical part of the oscillator design, since it now provides the additional high frequency feedback.

The oscillator circuit is shown in figure 5. The transistor is a Fairchild MT1050 which is in a coaxial package with very low lead inductances. The collector capacitance is of the order of 5 pf, the f_T is 1.5 GHz, and the dissipation rating is 2 watts. The tuning varactors are Varian VAT 54E. Two are used in series back to back to improve the capability of the circuit to accept a large rf voltage. The load coupling network is adjusted experimentally. Data is plotted in figure 6. At 1.4 GHz there is a change in slope of the tuning curve. This coincides with the point at which oscillation will cease if the capacitance C_i is removed. It appears that two modes of oscillation exist, depending upon which feedback mechanism is predominant. In order to account for the power output achieved, the rf voltage swing across the varactors must be quite large, and the analysis problem is a non-linear one. Second harmonic was measured to be 20 db down at 1 GHz and 35 db down at 1.2 GHz.

REFERENCES

1. J.F. Gibbons, "On the Analysis of the Modes of Operation of a Simple Transistor Oscillator", Proc. IRE, vol 49, pp 1383-1390, Sept. 1961.
2. K.M. Johnson, "Microwave Varactor Tuned Transistor Oscillator Design", IEEE Transactions on MTT, vol MTT-14, No. 11, Nov. 1966, pp 564-572.

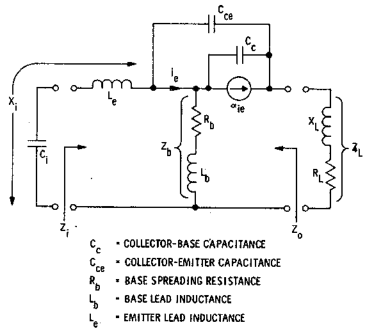

FIG. 1 - Transistor Oscillator Equivalent Circuit

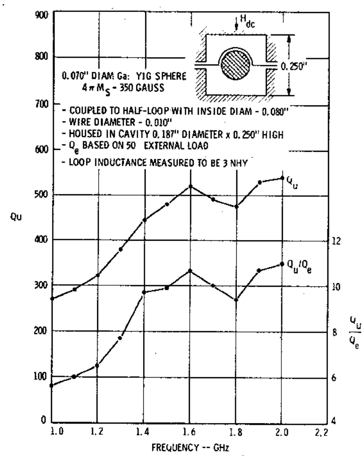

FIG. 2 - Measured YIG Resonator Q Data

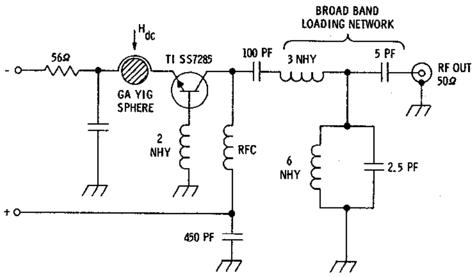

FIG. 3 - YIG Tuned Oscillator Circuit

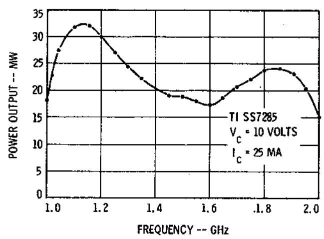

FIG. 4 - YIG Tuned Oscillator Power vs Frequency

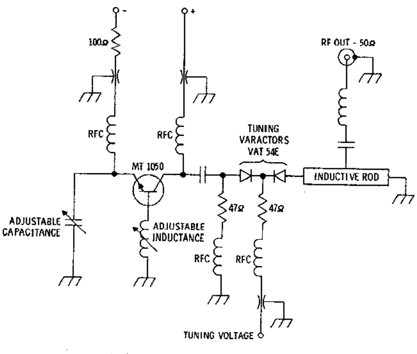

FIG. 5 - Varactor Tuned Oscillator Circuit

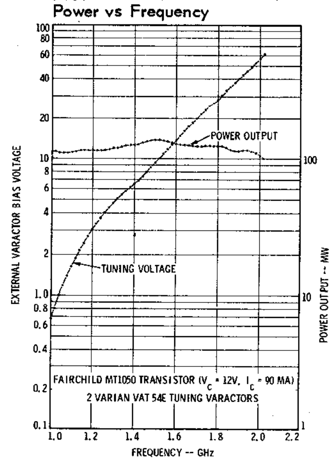

FIG. 6 - Varactor Tuned Oscillator Data