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A HYBRID GaAs MIC OSCILLATOR USING A MAGNETOSTATIC WAVE RESONATOR

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A Magnetostatic Surface Wave resonator and GaAs Field Effect transistor amplifier have been fabricated in an integrated format on a common MIC substrate operating at 3 GHz. The resultant oscillator had an output of -30 dB and phase noise of -92 dBc/Hz at 20 kHz offset.

Introduction

Magnetostatic Wave (MSW) Resonators [1,2] have exhibited the properties of high Q factors and wide tuning ranges. This paper describes the design and fabrication of a GaAs FET amplifier to drive an MSW resonator in a microwave oscillator. The devices were fabricated on hybrid thin film integrated circuits. The resonators used in this work have been fabricated on epitaxial Yttrium Iron Garnet (YIG) films grown on Gadolinium Gallium Garnet (GGG) [3]. MSW resonators using etched-groove gratings as reflectors have been reported [4] with loaded Q-factors of over 800. This same structure has been adopted in the present oscillator design. Gallium Arsenide Field Effect Transistors (GaAs FETs) have become an important component in solid state microwave amplifiers due to their low noise, high gain and high efficiency [5]. Distributed elements were used in the amplifier designs. Computer optimization of the circuit response was done using COMPACT [6]. The amplifiers were fabricated on 10 mil alumina substrates with gold metalization, to be compatible with resonator fabrication, which incorporated microstrips on the same kind of substrate. The resonator and hybrid MIC amplifier were assembled into the oscillator configuration of Figure 1.

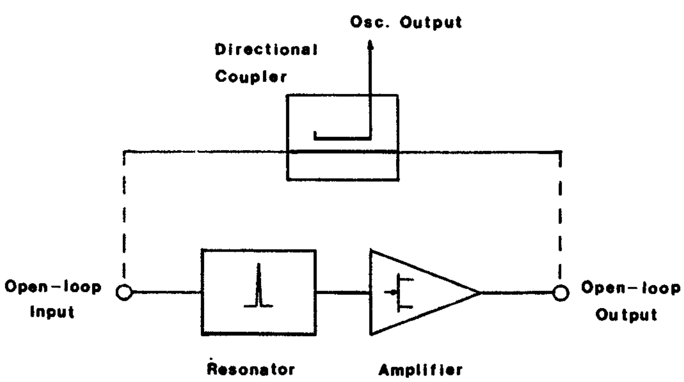

Fig. 1 Oscillator Block Diagram

Magnetostatic Wave Resonators

MSW Resonators offer the advantages of low-loss propagation and magnetic tunability that enable them to give excellent performance at microwave frequencies. The structure of the MSW resonator used consisted of etched groove arrays on a 20 um film of YIG, which was epitaxially grown on GGG and mounted on top of an alumina substrate with metal transducers. With the magnetic field applied parallel to the transducers, Surface Mode (MSSW) propagation is established [7]. The nonreciprocal characteristic of the surface wave helps to decouple the input and output ports when they are not in resonance. As the wave interacts with each reflecting element in the arrays, it is partially reflected. Overall reflection is maximum at a wavelength given by twice d, the period of the array. The frequency at which resonance occurs can be obtained by solving the dispersion relation of the MSSWs for the particular applied magnetic field, taking into account effects of the reflecting arrays. Since a change in the applied field simply shifts the dispersion relation along the frequency axis, the resonant frequency can be tuned by varying the applied field.

For the resonator used electrical signals were transduced to and from magnetostatic waves in the ferrite via hair-pin type transducers. This type of transducer was chosen for a better rejection of spurious responses at the low frequency end of the transducer pass-band [8]. The hair-pin pitch determines the resonator center wavelength. The transducer spacing determines the isolation of the direct electromagnetic feed-through. Transducer microstrip width and length combine to provide an optimum radiation resistance. The transducers were fabricated by etching gold plated substrates (IBE-M of Film Microelectronics, Burlington, MASS). The etched transmission line edges appeared to be ragged, as a result of etching 3 um of gold film. This led to some excess losses at microwave frequencies.

The dimensions of the resonator and transducers used are given in Table 1. The 3 dB bandwidth was 11

Table 1 Resonator Dimensions

Etched Groove Array		Transducers	
array spacing	3.6 mm	spacing	2.4 mm
groove width	72 um	pitch	150 um
land width	75 um	length	3 mm
groove depth	0.9 um	microstrip width	50 um

MHz, yielding a Q-factor of 270. Insertion and Return Loss were measured with the resonated peak tuned at different frequencies, by varying the applied magnetic field. The results are shown in Figure 2. Some spurious response was observed as the resonated peak was tuned across the band. These spurious responses are probably caused by self-resonance of the transducers. At 4.5 GHz and above, the spurious level

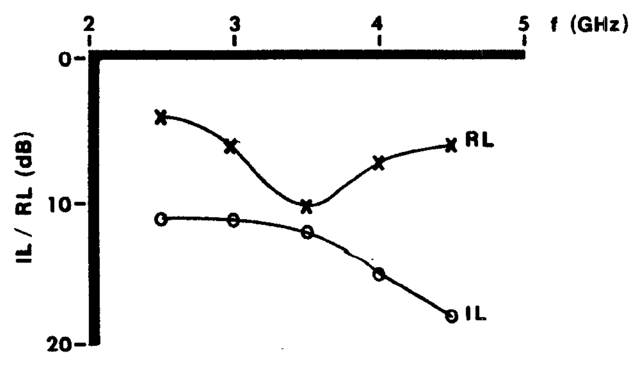

Fig. 2 IL and RL of the Modified Resonator

was higher than that of the resonated peak. In an oscillator feedback loop, this would cause oscillations at spurious frequencies when tuned to 4.5 GHz or above.

Single Stage FET Amplifier

An amplifier was designed to have a flat gain greater than 10 dB between 3 and 6 GHz. The Synthesis Method [9] was used in the design of matching circuits. In the initial design of the amplifier the FET was treated as unilateral, (S12 = 0), since the final circuit was computer optimized with the full set of experimental S-parameters of the transistor. The FET was matched to the signal source and load (generally 50 ohms) with lossless matching networks. The gain of the amplifier was treated as three distinct gain blocks [10]: The matching gain at the signal source, the gain of the FET, and the matching gain at the load. In order to obtain a flat gain over the entire bandwidth of interest, the sum of the three quantities (in dB) must remain the same at all frequencies. Representing the FET by its R-C equivalent circuit the matching problem can be reduced to that between a resistive source and load, with absorption of the capacitor into the matching circuit [11]. To keep the circuits simple, second-order Tchebyshev responses (requiring four essential elements) were used [12] to approximate the gain slope, though higher order networks can produce a slightly closer fit. The frequencies in the synthesis calculations were scaled to 6 GHz, the highest frequency in the band. Distributed instead of lumped elements were used in the amplifier implementation to eliminate parasitics that are associated with lumped elements. The amplifier circuit with distributed elements is shown in Figure 3 together with computer-optimized (COMPACT) element values in brackets.

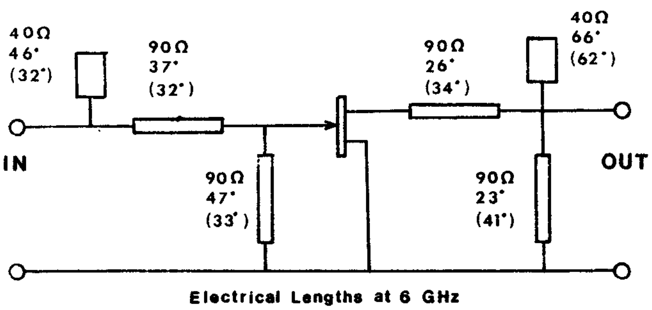

Fig. 3 Final Amplifier Circuit

The final circuit was implemented with microstrips on a 10 mil alumina substrate. High frequency isolation from the D. C. bias line was accomplished with high impedance (90 ohm) series quarter-wave transmission lines and ceramic chip shunting capacitors. Capacitors were also required at the D. C. input to suppress feedback at Megahertz frequencies. Alumina substrates with a thickness of 10 mils were coated with a thin layer of chromium (200A) and then gold (0.1um) without breaking vacuum. The gold would then be plated up to 3 um, forming the microstrip circuit. After plating, the photoresist was removed and the thin layers of gold and chromium were etched away, uncovering the microstrip circuit. The alumina substrate was mounted on an aluminum block. Holes were drilled on the substrate so that bypassing capacitors as well as the source leads of the transistor could be epoxied down directly on the aluminum ground block. The chip capacitors were wire-bonded to the microstrips with 2-mil gold wires. The SMA connectors were screwmounted with the center tabs bent into contact with the microstrips and silver epoxy was then applied.

Single Transistor Oscillator

The matched resonator described was cascaded with a single-stage FET amplifier. Oscillation was obtained by feeding the output from the amplifier back to the resonator, as shown in Figure 1. The amplifier consisted of an FET matched to input and output impedances of 50 ohms using transmission line sections. The design was narrow-banded, to ensure enough gain for oscillations. The layout of the resonator-amplifier cascade is shown in Figure 4, in which the resonator was matched to 50 ohms via

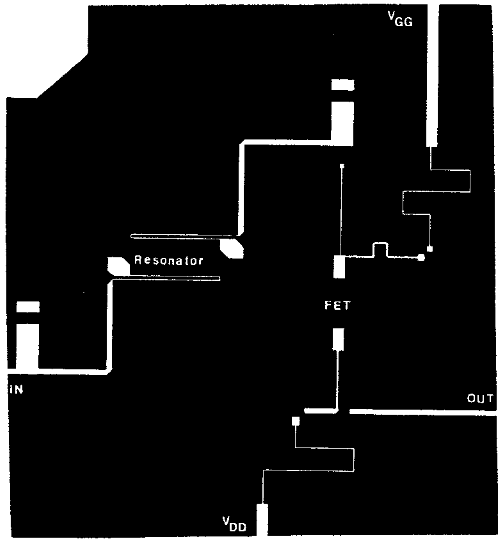

Fig. 4 Layout of the Oscillator

a transmission line and a capacitive stub. Output to the amplifier was fed through a DC-blocking capacitor. High impedance microstrips and ceramic capacitors were used for decoupling the biasing circuit. The output 50-ohm lines were bonded to another DC-blocking capacitor. The circuit was etched on a gold-plated alumina substrate.

In the closed loop configurations, an output of -30 dBm was observed at the coupling port of the -10 dB directional coupler. The oscillation frequency was tunable from 3.08 to 3.19 GHz. The YIG-resonator used in this oscillator was found to saturate at a power level of -20 dBm, thus limiting the output to -30 dBm at the coupler.

Phase Noise for resonator stabilized oscillators is given by Lewis [13]:

$$L(f) = 10 \log \frac{GkTFf^2}{Q^2 P_o (f-f_o)^2} \quad dB_{c/Hz}$$

where G is the amplifier gain, k is Boltzmann's constant, T ambient temperature (K), F the noise figure of the amplifier, f_o the carrier frequency, Q the quality factor of the resonator, and P_o the loop power.

The oscillator phase noise was measured with an HP8566 Automatic Spectrum Analyser System and the result is shown in Figure 5, together with the computed values using the amplifier noise figure as a parameter. For the calculated values, Q = 270, G = 63 (18 dB), f_0 = 3.1 GHz, and P_0 = 0.01 mW (-20 dBm). The measured result is very close to the predicted one, although slightly higher, probably due to a higher Q value, and a lower actual gain value at 3.1 GHz as compared to 3.0 GHz.

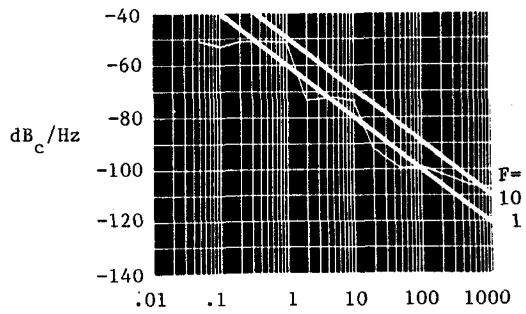

Fig. 5 SSB Phase Noise at 3.135 GHz

Conclusion

A resonator and an amplifier have been integrated onto one MIC substrate. With the inclusion of the -10 dB coupler a truly integrated oscillator can be achieved. The output power can be increased by improving the quality of the YIG resonator to raise the saturation power level. To extend the sweeping range, the transducers need to be optimized to reduce the insertion loss, and broadband matching circuits will be required.

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