A BROADBAND VCO USING DIELECTRIC RESONATORS

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ABSTRACT

A novel approach for obtaining broadband negative resistance in designing 1-port oscillators is described. The circuit uses a varactor in the feedback path to obtain a variable reactance for maximizing the negative resistance. The concept is demonstrated by designing and fabricating a circuit that oscillates in the 3.5- to 6.5-GHz band by only replacing the dielectric resonator. The technique will be useful in building broadband sources and also in MMIC technology as a tool for electrically tweaking the circuits.

INTRODUCTION

Dielectric resonator oscillators are increasingly being used as the local oscillators in the telecommunications systems. Typically, dielectric resonator oscillators are designed and built to oscillate at a single frequency of operation. The approach described enables the circuit to oscillate through a broad range of frequencies limited only by the gain of the transistor and the capacitive swing available from the varactor. The goal was to design a voltage controlled oscillator for the 3.7-4.2 GHz and 5.8-6.3 GHz common carrier frequency bands. The design uses a common circuit for the 3.5-6.5 GHz band and needs only the dielectric resonator to be replaced to set the oscillator to the desired frequency. Electronic tuning for AFC is provided by a tuning varactor coupled to the tank circuit. The VCO was used in a prototype broadband down converter for a 4-6 GHz microwave digital radio.

CIRCUIT DESIGN

A bipolar transistor, NEC 21900, is used for low phase noise compared to a GaAs FET. Initial design is similar in approach to the design of any oscillator. An oscillator circuit can be divided [1] into an active 1-port $Z_D(A)$ and a tank circuit $Z_L(w)$ as shown in Fig. 1.

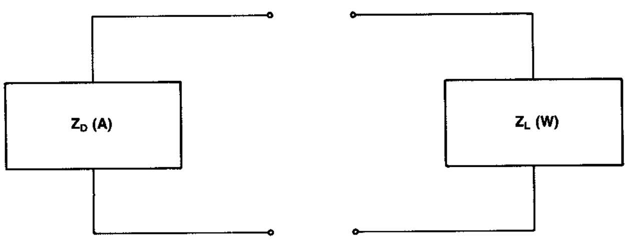

Figure 1. Active One Port and Tank Circuit for the Oscillator.

For steady state oscillations

$$Z_D(A) + Z_L(w) = 0$$

Therefore, a broadband negative impedance $Z_D(A)$ originating from the active device and a tank circuit which will determine the frequency are needed. The above equation then leads to the frequency w and the amplitude A of the oscillations.

This analysis is based on the active device being a 1-port network. The transistor, being a 2-port network, has to be converted first to a 1-port network. After comparing the possible configurations analytically, it was concluded that the common collector configuration with an inductive reactance at the base yielded the maximum negative resistance $Z_D(A)$. The result of mapping the base plane into the emitter plane at 5 GHz indicated that j42 ohms or 1.3 nH is required at the base for the maximum reflection coefficient. This can be obtained by bonding a length of wire equal to the inductance desired (80 mil for 1.3 nH) at the base. Fig. 2 shows the magnitude of S_{11} plotted against the frequency with different feedback inductance values. It can be seen that at L = 1.3 nH, S_{11} is greater than 1 between 2.5 and 6.5 GHz with S_{11} peaking to >4 at 5 GHz as expected. With proper tank circuit, this circuit will produce maximum power at 5 GHz. At other frequencies, however, the magnitude of the negative resistance may not be large enough to start oscillations or produce maximum power. The feedback inductance will therefore have to be adjusted to produce maximum negative resistance at different frequencies.

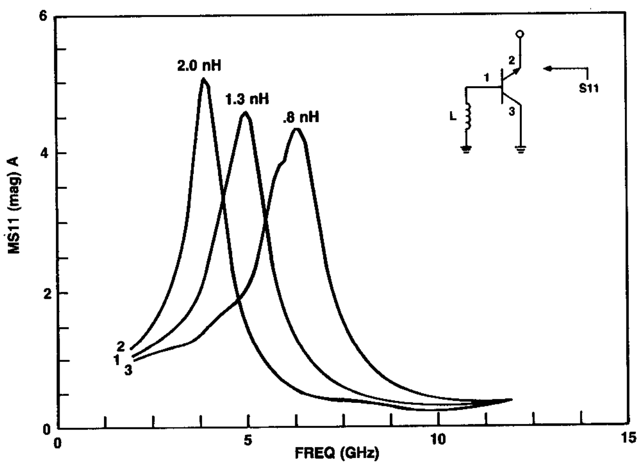

Figure 2. Calculated S_{11} versus Frequency.

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A circuit to obtain this is shown in Fig. 3. It uses a varactor chip, M/A 46574D, and a bypass capacitor in series with the inductor. The series combination above resonance acts like an inductor. As the voltage across the varactor is changed, capacitance changes, and so the effective inductance at the base changes. This will enable peaking of the reactance for maximum negative resistance at the desired frequency and thus maximize power. Fig. 3 shows the calculated S_{11} versus frequency with 2.5 nH inductor and a varactor with a capacitive swing from 5 pF to 0.35 pF at the base. It is evident that a very broadband negative resistance can be obtained by this method limited only by the available capacitive swing from the varactor.

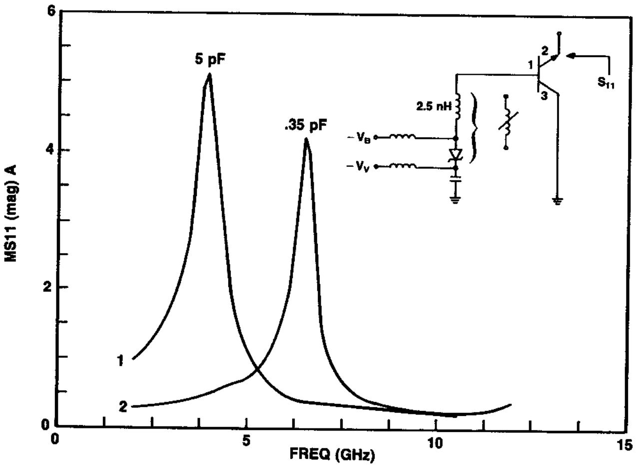


Figure 3. Calculated S_{11} versus Frequency at Two Varactor Capacitance Values.

This analysis is experimentally verified using the device line method [2]. The transistor and the varactor chip are mounted inside a housing using gold epoxy. The emitter pad of the transistor is bonded to a 50 ohm microstrip transmission line. An RT/duroid 6010.2 is used for the microstrip line. At different frequencies, the input power level is increased until the added power becomes maximum. At the same time, voltage on the varactor is adjusted to maximize the added power. Fig. 4 shows the maximum added power

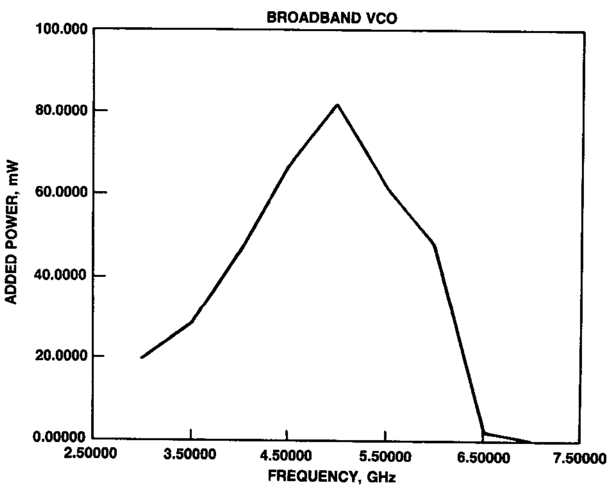


Figure 4. Maximum Added Power versus Frequency.

versus the frequency under above conditions. It is evident that this circuit is very unstable from 3.0 to 6.5 GHz and will oscillate in this frequency range with a proper tank circuit.

OSCILLATOR CIRCUIT

Having achieved a broadband negative resistance looking from the emitter, the next step is to provide an appropriate load for the desired oscillations. This is done through the use of dielectric resonators at different frequencies. The dielectric resonator is placed near the 50-ohm line as shown in Fig. 5 and essentially operates as a transmission type resonator [3] between the active device and the outside load. The dielectric resonator is selected for the desired frequency and placed on the line for the circuit to oscillate. A quartz spacer of 80 mil height is used to support the dielectric resonator. Fig. 6 shows the photographs of the same circuit at 4- and 6-GHz using 4 and 6 GHz dielectric resonators respectively. It was found that the placing of the dielectric resonators in the circuit was not critical.

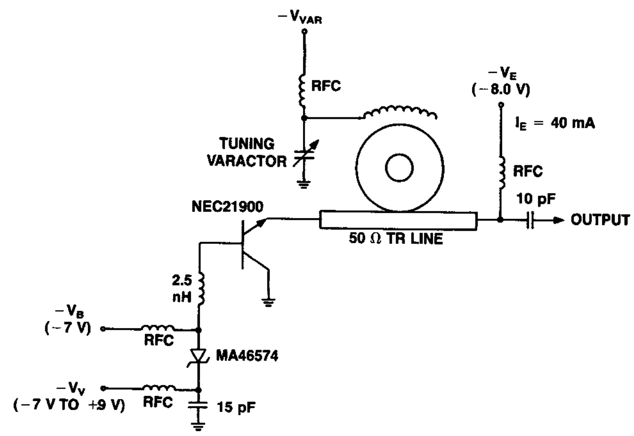


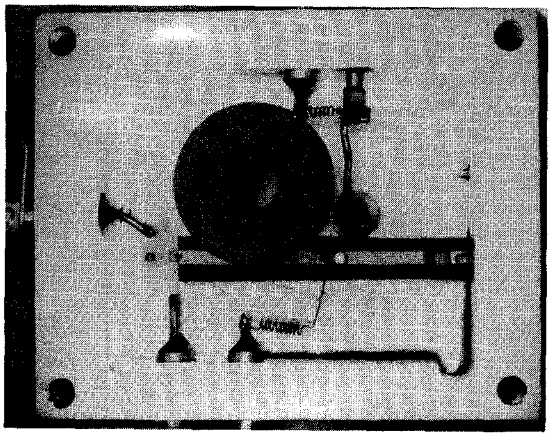
Figure 5. Dielectric Resonator Oscillator Circuit.

A lumped circuit model [4] at 4 GHz is used for the dielectric resonator for the analysis of the oscillator as shown in Fig. 7. It also shows the calculated magnitude of S_{11} plotted against the frequency. It can be seen that mag $[S_{11}]>100$ at 4 GHz, indicating oscillations at 4 GHz. Fig. 8 shows the output power in dBm when dielectric resonators of different frequencies are placed in the circuit and the voltage at the varactor is adjusted to maximize the power. It can be seen that the oscillator produces a +13 dBm minimum power for a frequency range of 3.9 to 6.7 GHz. There is a good correlation between this and the curve in Fig. 4. There is no data between 3.5 and 3.9 GHz and 4.5 and 6.4 GHz as the dielectric resonators were not available, but it can be safely concluded that the circuit will oscillate anywhere in between.

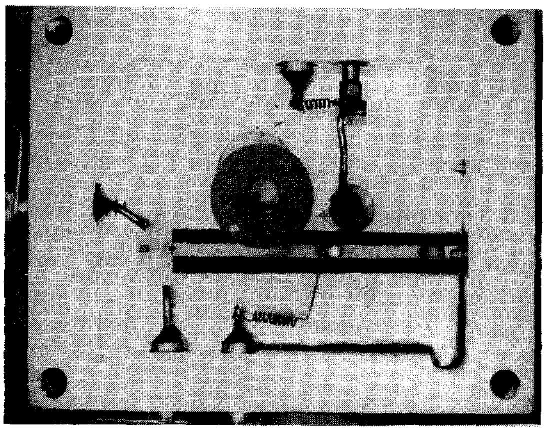
Typical biasing is shown in Fig. 5. To cover the 3.5-6.5 GHz frequency range, the varactor voltage has to be changed from -7.0 to +9.0 volts, which corresponds to reverse biasing the varactor from 0 to 16 volts. It turned out that the negative voltages at the varactor terminal covered the 4 GHz band while positive voltages covered the 6 GHz band.

TUNING

The oscillator is mechanically tuned using a tuning screw above the dielectric resonator. A mechanical tuning range of 80 MHz minimum



(a) 4.0 GHz



(b) 6.0 GHz

Figure 6. Photographs of the Oscillator.

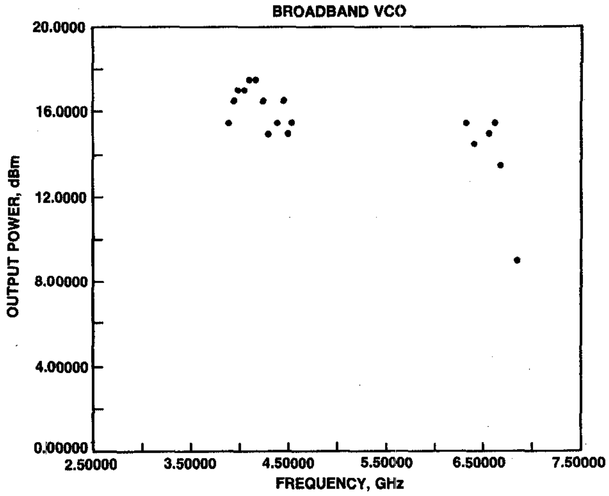


Figure 8. Output Power from the Oscillator at Different Frequencies.

is obtained at 4 and 6 GHz. Electronic tuning is achieved by inductively coupling a varactor to the dielectric resonator as shown in the photographs in Fig. 6. The tuning inductor is supported on a quartz spacer to reduce the microphonic effects. The data for electronic tuning was taken at five different frequencies to cover the whole band. Fig. 9 shows the frequency and power variation respectively with the tuning voltage. The tuning rate varies because of the change in coupling between the varactor and the dielectric resonator at different frequencies. If uniform tuning is needed, it is very easy to obtain it by simply adjusting the gap between the inductive wire and the dielectric resonator.

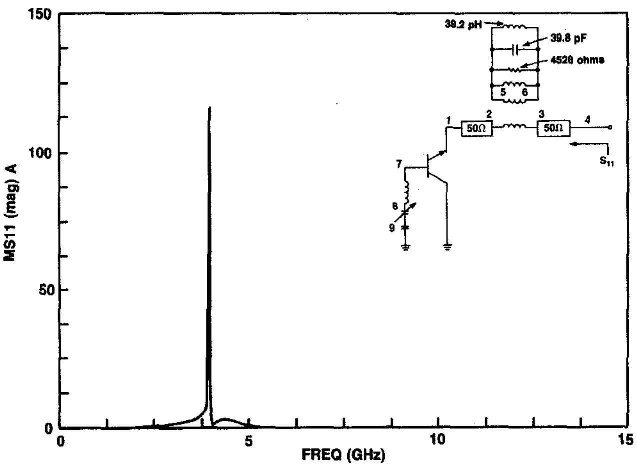


Figure 7. Calculated S_{11} versus Frequency for the Dielectric Resonator Oscillator.

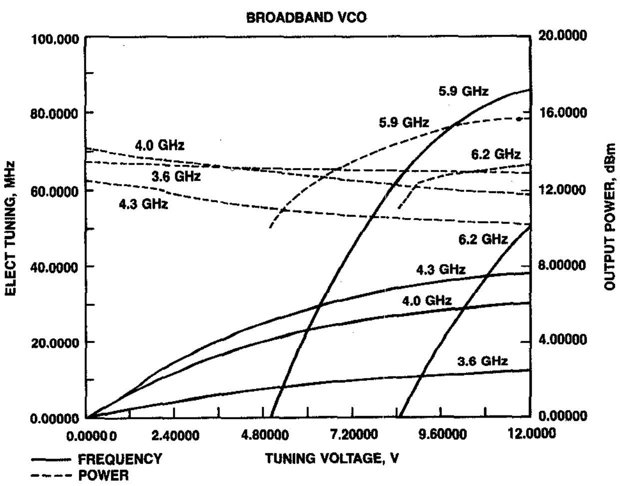


Figure 9. Electronic Tuning at Different Oscillator Frequencies.

PHASE NOISE AND TEMPERATURE EFFECTS

A typical phase noise curve for the oscillator at 6 GHz is shown in Fig. 10. The close-in phase noise at 1 kHz offset from the carrier is -75 dBc/Hz, which is typical of the oscillators using dielectric resonators. The frequency stability and power variation with temperature is shown in Fig. 11 at 4 and 6 GHz. The maximum frequency drift is 4 MHz for operation over the temperature range of -30 to 70°C. The oscillator was used in a phase-locked circuit which easily compensated for the temperature drift.

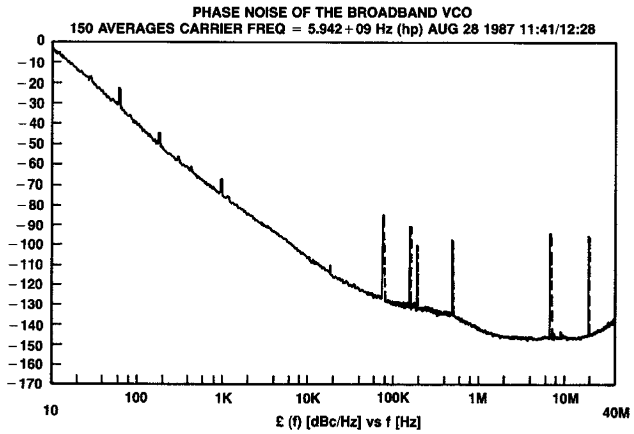


Figure 10. Phase Noise Characteristics.

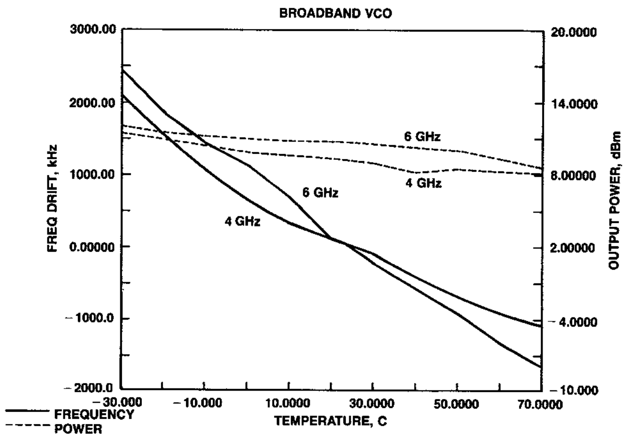


Figure 11. Frequency Stability with Temperature.

CONCLUSION

A simple approach for designing a broadband dielectric resonator oscillator has been described. The circuit uses a bipolar transistor with a series resonant feedback as the active device and a dielectric resonator as the load. The circuit oscillates in the 3.5 to 6.5 GHz band by simply replacing the dielectric resonator. The oscillator is mechanically tuned using a tuning screw and electronically tuned using a varactor diode. Four oscillators were built and tested successfully to ensure consistency and feasibility. The VCO was used as part of a 4-6 GHz phase-locked oscillator in a microwave digital radio.

ACKNOWLEDGMENT

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