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Using Hewlett-Packard MSA Series MMIC Amplifiers as Frequency Converters

Application Note S005

Introduction

Hewlett-Packard's MSA Series amplifiers can be used as the active element for an oscillator at frequencies up to 8 GHz or as a mixer with gain for RF input frequencies up to 9 GHz or as a self-contained frequency converter (self-oscillating mixer). As a mixer, these devices can be used as either upconverters or downconverters while providing conversion gain for RF signals up to 9 GHz. Figure 1 shows typical conversion gain vs. frequency for two models configured for active frequency conversion along with the associated functional RF schematic and block diagram. MSA series devices are available in a variety of metal, ceramic and low-cost plastic stripline packages for military, commercial and consumer applications.

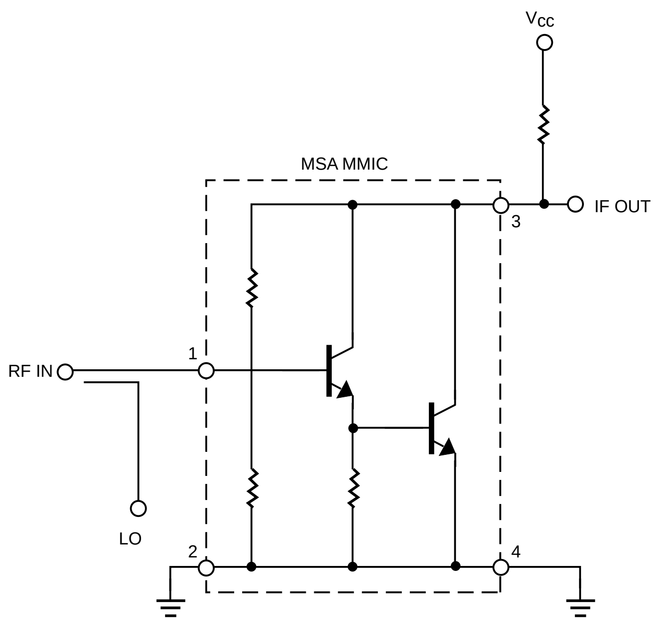


Figure 1a. Functional RF Schematic for Active Frequency Conversion. (The normal test configuration for MSA series devices uses an external LO as shown here. The normal operating configuration for self-oscillation is shown in Figure 2.)

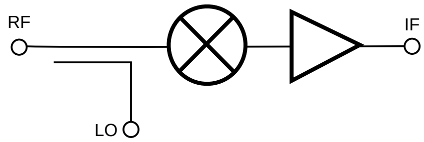


Figure 1b. Functional Block Diagram for Active Frequency Conversion.

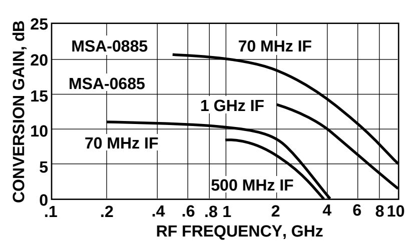


Figure 1c. Typical Conversion Gain vs. Frequency.

The monolithic microwave integrated circuit (MMIC) used in the frequency converter consists of a silicon bipolar Darlington pair with associated bias resistors. It is fabricated using a 10 GHz f_t nitride self-aligning process featuring ion implantation and thin film poly-silicon resistors. The geometry incorporates inter-digitated sub-micron-width emitter fingers with an emitter pitch of 4 microns, the centerline distance between adjacent emitter fingers. The circuit schematic for the frequency converter Darlington pair is shown within the dotted lines in Figure 1a.

Theory

The basic feedback configuration used to achieve oscillation is shown in Figure 2. For the circuit to oscillate, the amplitude and phase criteria must be satisfied, that is $A\beta = 1$. If an RF signal is also applied to the input of the frequency converter, it will mix with this local oscillator signal, producing sum and difference products at the output. Transistor Q1 of the Darlington pair (Fig. 1) is the non-linear element that limits the amplitude of the oscillation and generates the frequency products. The second stage (Q2) provides IF amplification. Since the self-oscillating mixer is unbalanced, filters must be added to the input and output of the device if the generated frequency products are to be limited to the desired sum or difference frequencies.

The feedback network can be realized in a number of ways. Two of the most common tank circuits are the dielectric resonator, which has a resonant frequency determined by its physical properties; and a simple parallel or series lumped inductor and capacitor, which has a resonant frequency determined by the equation:

$$f_r = \frac{1}{2\pi\sqrt{LC}}$$

In either case the phase shift through the feedback path and the device must be 360 degrees at the frequency of oscillation. At frequencies below approximately 3 GHz, the dimensions of available dielectric resonators become impractically large, and lumped-element resonant circuits or surface-acoustic wave structures are more suitable.

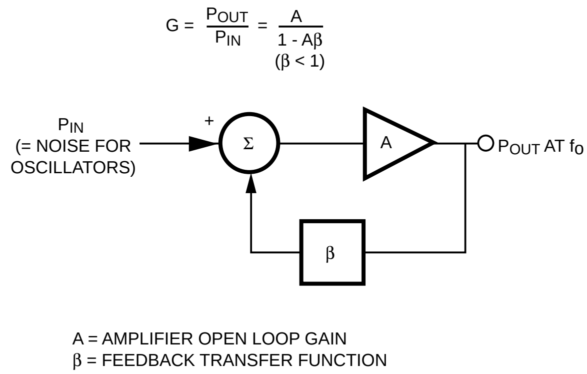


Figure 2. Block Diagram of the Basic Feedback Configuration to Achieve Oscillation.

Dielectric Resonator Design Considerations

The dielectric resonator is made of a low loss, temperature stable, high permittivity and high Q ceramic material in a regular geometrical form. It resonates in various modes at frequencies determined by its dimensions and shielding conditions. Because of its small size, low price and excellent integrability in MICs, its application in active and passive microwave circuits has been rapidly increasing. For example, barium titanate has a relative dielectric constant (ε_r) of 39, an unloaded Q of approximately 7000 at 6 GHz, and a temperature stability of approximately 2 ppm/°C. Other dielectric materials are also available[1,2], including materials with similar dielectric characteristics but with a selection of temperature coefficients, permitting the selection of a resonator to compensate for the oscillator temperature-induced frequency drift caused by the change in active device parameters and the substrate expansion. See Reference 3, Plourde and Ren, for a thorough presentation of this topic.

If a signal can be coupled into the dielectric resonator by exciting its external magnetic field, for example through coupling to a transmission line or by placing the resonator in a waveguide, the dielectric resonator acts as a dual of a high-Q resonant cavity in the sense that electrical short-circuit boundaries (metallic walls) are replaced by magnetic short-circuit boundaries (dielectric walls). Although both E and H modes can be excited in a dielectric resonator, the $TE_{01\delta}$ mode has a large normal magnetic field component at the boundary surface, making it an easily-usable mode.

The dimensions and surroundings of a dielectric resonator determine its resonant frequency. A common configuration used in MICs is shown in Figure 3 where the dielectric resonator is placed on a substrate in a metallic shielding box which affects the resonant frequency and unloaded Q factor of the resonator. In this configuration, the resonant frequency can be accurately determined by methods using high-level computers. Kajfez[4] presents an approximate solution which allows the use of desk top computers in determining the resonant frequency with an accuracy of better than ±2%. An aspect ratio D/H of 2.2 to 2.8 is

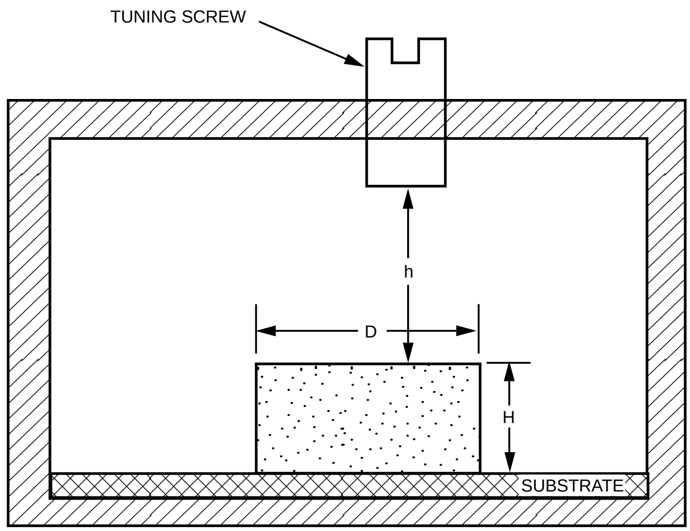


Figure 3. Dielectric Resonator on a Microstrip Substrate.

used for a good quality factor and optimum separation of higher order modes.

Environmental conditions, such as the close proximity of conducting walls or dielectric materials affect both the resonant frequency and Q of the dielectric resonator. Fortunately, the effects on resonant frequency of the presence of the conductor plane of a microstrip substrate are usually very small. If the thickness of the substrate is less than one-quarter of the puck thickness, and the dielectric constant of the substrate is less than one-half that of the puck, the shift in frequency due to the substrate is less than 1%. If structural walls are placed at a distance greater than the puck diameter, their effect should also be negligible[5].

Design Example: C-Band TVRO Downconverter with dielectric resonator using MSA-0870.

In designing a frequency converter using a dielectric resonator, the first step is to select a resonator operating at the required LO frequency, with the required stability characteristics. Knowing the dimensions and characteristics of the resonator, it becomes possible to develop a circuit layout that provides the appropriate degree of coupling to the resonator.

Our example is a downconverter suitable for television receive-only (TVRO) earth stations[6], using a high-side LO frequency of 5.15 GHz. The RF input is 3.7 GHz to 4.2 GHz, and the IF band 1.45 GHz to 0.95 GHz.

An RF schematic of this circuit is shown in Figure 4. The dielectric resonator is coupled between the input and output of the MSA-0870, with the coupling coefficient controlling the overall conversion gain and output 1 dB compression point.

In a practical layout (Fig. 5), the DR is placed at an arbitrary distance from each transmission line (10 mils for example), with the coupling optimized by varying the spacing between the DR and the substrate dielectric material for optimum performance. This is accomplished by substituting low-loss spacers of different thicknesses (ranging from a few mils to approximately 100 mils). The length l_1 should be 90 degrees

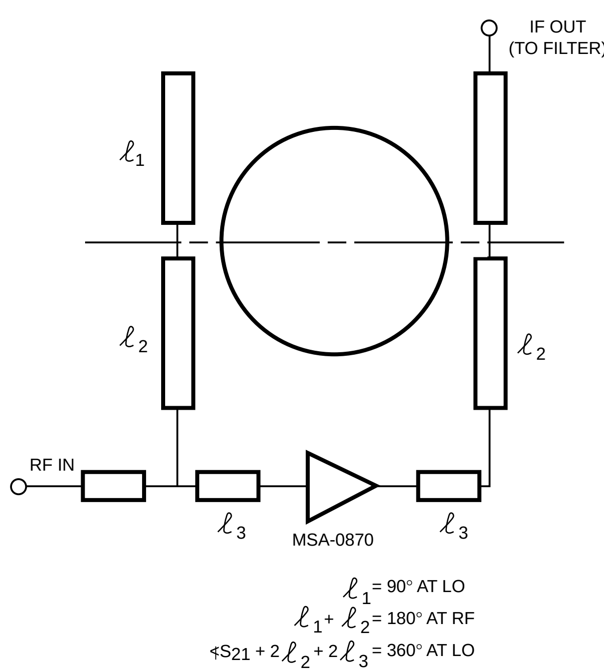
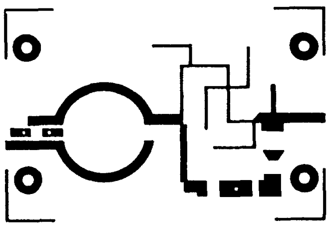


Figure 4. RF Schematic of the C-band, TVRO Downconverter.



COMPONENT COUNT:
1 Si MSA
1 PC BOARD
2 RESISTORS
1 CAPACITOR
1 DIELECTRIC PUCK
1 RF CHOKE* (OPTIONAL)
2 OFF BOARD BLOCKING CAPACITORS

Figure 5. Actual Size PC Board including Output low-pass Filter.

```
CKT
        MSUB   ER=4.9  H=31  T=.2  RHO=1  RGH=0
        MLIN   4  1  W=12  L=300
        MLEF   1  W=12  L=315
        MLIN   1   2   W=11   L=315
        MLEF   2   W=17   L=312
        MLEF   2   W=17   L=312
        MLIN   2   3   W=11   L=315
        MLEF   3   W=12   L=315
        MLEF   3   W=20   L=150
        DEF2P   4   3   FIL

FREQ
        SWEEP    .5    12    .2

OUT
        FIL    DB[S21]   GR1

GRID
        RANGE   0   12   1
        GR1   -30   0   5
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Figure 6. Touchstone™ Circuit File for the Band-stop Filter.

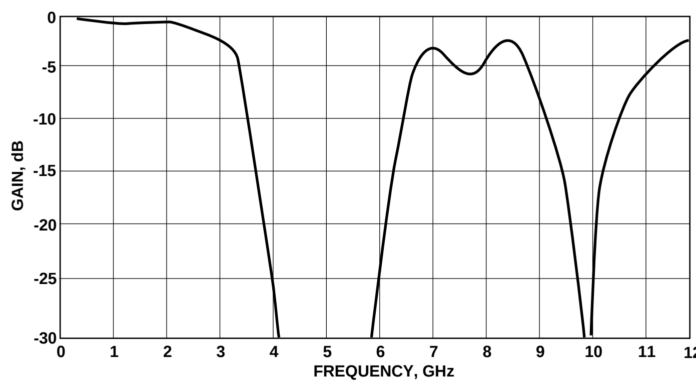

Figure 7. Filter Response Curve Generated by the Program in Figure 6.

at the frequency of oscillation. This will reflect a low impedance, hence a current maximum, at the point where the DR is coupled to the transmission line. The lengths ($l_1 + l_2$) should be equal to 180 electrical degrees at the RF frequency, which will reflect a high impedance at the RF input, to prevent loading. The lengths ($2 \times l_2 + 2 \times l_3$) plus the phase shift through the device should be 360 degrees at the frequency of oscillation. The 180° phase shift through the DR provides the correct current direction for the oscillator. With the above criteria satisfied the MSA series device will oscillate at 5.15 GHz.

Filter Design

A band stop filter is designed to reject the RF and LO signals present at the output. A three-section band-stop Chebyshev filter is implemented which exhibits 0.5 dB of ripple, and 30 dB of rejection at the LO frequency. An additional quarter-wave open circuit stub, to reject the 2LO frequency of 10.3 GHz is also added. The *Touchstone*™ circuit file for this filter is given in Figure 6, and the response curve generated by the program is shown in Figure 7.

Biasing

The MSA series product is a current-controlled device, and its bias point can best be specified by its total device current. The MSA-0870 is specified with a normal operating current of 35 mA at a nominal device voltage of 7.5 volts. A simple bias stabilizing resistor should be inserted in the output feed, with the proper RF choke and power supply decoupling capacitors. This will act as a feedback element and help stabilize the device current. A four volt drop is sufficient to achieve effective compensation. A more complete discussion of DC biasing circuits is given in Avantek Application Note AN-S003[7].

Experimental Results

A prototype has been built using the MSA-0870. The circuit (Fig. 5) is fabricated on 31 mil thick, epoxy-glass (FR4) board ($\varepsilon_r = 4.8$). The photo of Figure 8 shows how the MSA series device and dielectric puck are mounted on the board.

The prototype downconverter exhibits a typical conversion gain of 9 dB, a SSB noise figure of 13 dB, input and output VSWR of less than 2.5:1, 1 dB compression

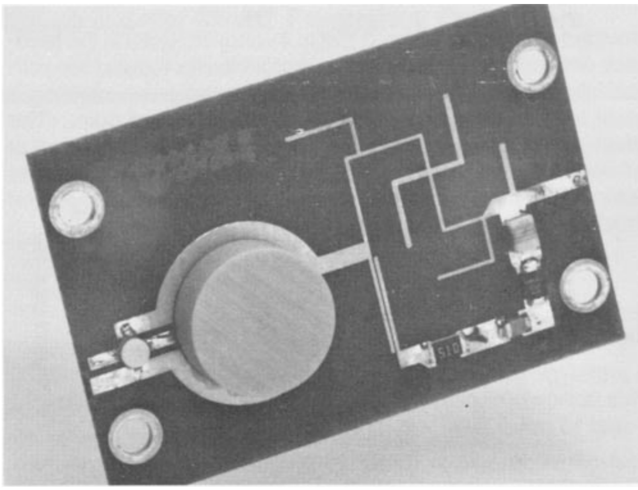

Figure 8. TVRO Circuit Board with Dielectric Puck.

point of +8 dBm, and third order intercept of +17 dBm. A significant LO frequency product will be present at the RF and IF ports, requiring external filtering.

Design Example: 1.575 GHz GPS downconverter with lumped LC tank circuit using MSA-0670

For a GPS receiver, the RF input frequency of 1.575 GHz is down-converted to an IF of 0.175 GHz. The required 1.4 GHz low side LO frequency is rather low for the use of a dielectric resonator as the frequency-determining element. However, the tank circuit for the self-oscillating mixer can easily be produced by using a series RLC lumped-element network.

The schematic for the GPS downconverter is shown in Figure 9, with the series RLC network (tank circuit) inserted between the input and output. The resonant frequency of the series circuit, and the approximate frequency of oscillation is given by the equation:

$$f_r = \frac{1}{2\pi\sqrt{LC}}$$

A transmission line segment in series with the device insures that the total electrical length through the device and the tank circuit is equal to 360 electrical degrees. The 20-ohm resistor will determine the loss of the tank, and by varying its value the converter can be optimized for maximum gain and output 1 dB compression point.

The feedback network should be designed so that at the input it looks like a high impedance at the RF frequency and at the output like a high impedance at the IF frequency. Using this technique the designer will essentially decouple the tank circuit from the frequency converter.

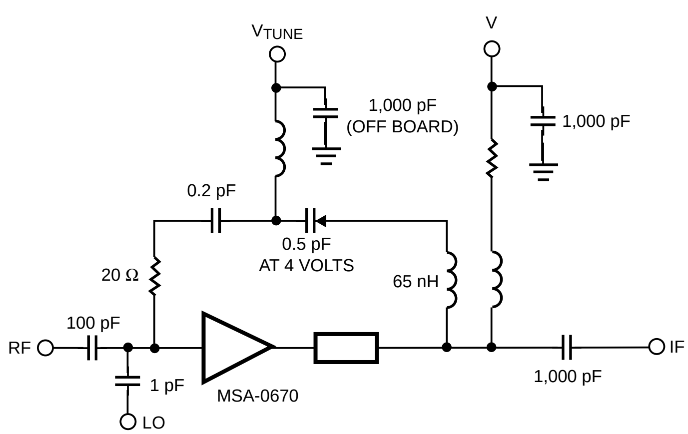

Figure 9. Schematic for the Prototype GPS Downconverter.

Two special features are incorporated in this schematic. Some degree of voltage-controlled fine tuning (a few hundred MHz) is provided with the varactor (0.5 pF at 4 volts) included in the feedback network. The LO input circuit provides for locking the self-oscillating mixer to a stable frequency source, by injecting a small amount of reference signal at the indicated point. (The advantage of this technique is that only a very small amount of reference signal power is required.) This injection capacitor reduces the conversion gain in the demonstration circuit, and should be omitted if it is not needed.

Experimental Results

A prototype downconverter has been built using the MSA-0670. This device is chosen for its lower power consumption, only 3.5 V and 16 mA.

The circuit shown in Figure 10 is fabricated on 25 mil thick, Epsilon 10TM board ($\varepsilon_r = 10$). This photo (Fig. 10) shows how the MSA series device and components are mounted on the board.

This downconverter exhibits a typical conversion gain of 9 dB, a

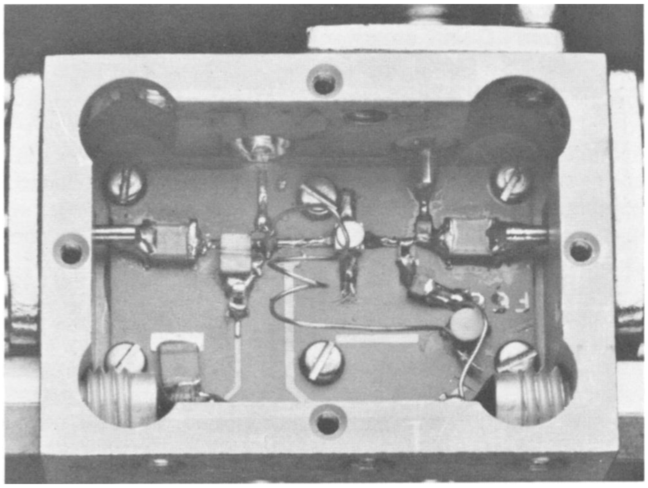

Figure 10. GPS Downconverter Circuit Board.

SSB noise figure of 15 dB, input and output VSWR of less than 2:1, 1 dB compression point of -2.5 dBm, and third order intercept point of +7.5 dBm. Again a significant LO product will be present at the RF and IF ports, which should be limited by filtering the input and output or by using a preamplifier on the input.

Conclusion

This application note has presented the concept of frequency conversion using a silicon MMIC with conversion gain and has illustrated two simple applications. HP MSA series amplifiers are ideally suited for very low cost or size constrained applications in such markets as CATV converters, satellite MATV and TVRO block converters, downconverters in UHF/VHF communications systems, 2nd IF converters in police radar detectors, cellular radio receivers, MDS television receivers, GPS and INMARSAT receivers and military manpack systems. There are literally thousands of potential applications where adequate conversion gain flatness and isolation can be achieved using simple external filters.

References:

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2. Trans-Tech, Inc., 5520 Adamstown Road, Adamstown, MD 21701 .
3. J.K. Plourde and C.L. Ren, “Application of Dielectric Resonators in Microwave Components,” *IEEE Trans. MTT*, Aug., 1981 pp. 754-770.
4. D. Kajfez, “Elementary functions procedure simplifies dielectric resonators’ design,” *Microwave Systems News*, June 1982, vol. 12, pp. 133-140.
5. R.R. Bonetti and A.E. Atia, “Analysis of Microstrip Circuits Coupled to Dielectric Resonators,” *IEEE Trans. MTT* December, 1981, pp. 1333-1337.
6. I. Kipnis and A.P.S. Khanna, “10 GHz Frequency-Convertor Silicon Bipolar MMIC,” *Electronics Letters*, November 1986, Vol. 22, pp. 1270-1271.
7. Application Note AN-S003. “Biasing MODAMP™ MMICs”.

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Obsoletes 5963-3649E

Printed in U.S.A. 5965-7391E (3/97)