Microcircuits for the Microwave Sweeper

By Ronald E. Pratt, Robert W. Austin and
Arlen Dethlefsen

VIRTUALLY ALL THE RF CIRCUITRY in the HP Model 8620A Microwave Sweep Generator is hybrid thin-film microcircuitry developed and fabricated in house. The frequencies from 0.1 to 4.2 GHz are produced by two rf modules which plug into the sweeper's rf drawer. One of these modules (Model 8633A) has a YIG-tuned transistor oscillator that is swept from 1.8 to 4.2 GHz. The second module (Model 8632A) heterodynes with the first to produce a swept output from 0.1 to 2.0 GHz. Fig. 1 is a block diagram of the microcircuits used to generate the 0.1 to 4.2 GHz swept signal.

Microcircuits for the 8632A Heterodyne Module

The design objective for the 8632A Heterodyne Module was to provide a 0.1-to-2.0 GHz signal with a power level in excess of +10 dBm, with spurious and harmonic signals more than 30 dB down from the fundamental, internal leveling, and a minimum of space and complexity. These were the techniques used to accomplish this:

- Heterodyning a 2.3 to 4.2 GHz YIG oscillator with a 4.3 GHz local oscillator allowed coverage of the 0.1 to 2.0 GHz band in one continuous sweep.
- The YIG oscillator signal was used as the low level signal. This eliminated the need for switching the position of the modulator or the use of two modulators for the two frequency bands.
- Troublesome spurious signals and local oscillator leakage were minimized by using a local oscillator frequency of 4.3 GHz. This choice of LO frequency also allows the swept signal to be used below 100 MHz since the two mixing signals (4.2 and 4.3 GHz) are easily filtered out.
- A PIN transfer switch eliminated the need for a bulky, possibly expensive mechanical switch, its associated rf cabling and limited life.
- A resistive bridge type directional detector accomplished internal leveling with a minimum of space and cost.
- A high-gain wideband amplifier minimized the spurious signal levels and provides the required output power.

The 8632A Heterodyne Module contains two microcircuit packages, a frequency converter package and a 0.1–2 GHz amplifier package.

0.1–2.0 GHz Frequency Converter

The frequency converter contains five separate microcircuit substrates:

1) Directional Detector/Switch
2) Filter/Coupler
3) 4.3 GHz Oscillator
4) Mixer
5) 2 GHz Low Pass Filter

Thin film microcircuit techniques have made it possible to house these circuits in a 2 x 2 x ½ inch package.

Directional Detector/Switch

Fig. 2 is a simplified schematic of the circuitry of the Directional Detector/Switch substrate and Fig. 3 shows the circuit layout. The circuitry is contained on a 1 x 0.25 x 0.025 inch sapphire substrate. The directional detector portion of the circuitry is contained on the left side of the figure and the transfer switch circuitry is contained on the right.

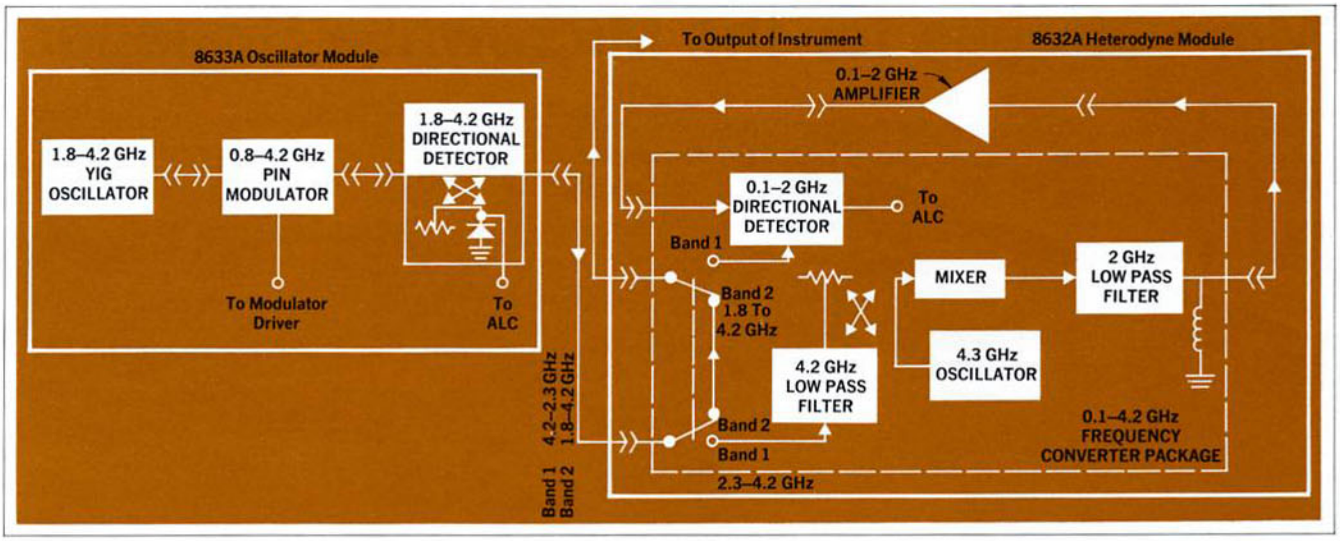


Fig. 1. *Block diagram of microcircuits used in 8620A Sweep Signal Generator to produce signals from 0.1 to 4.2 GHz.*

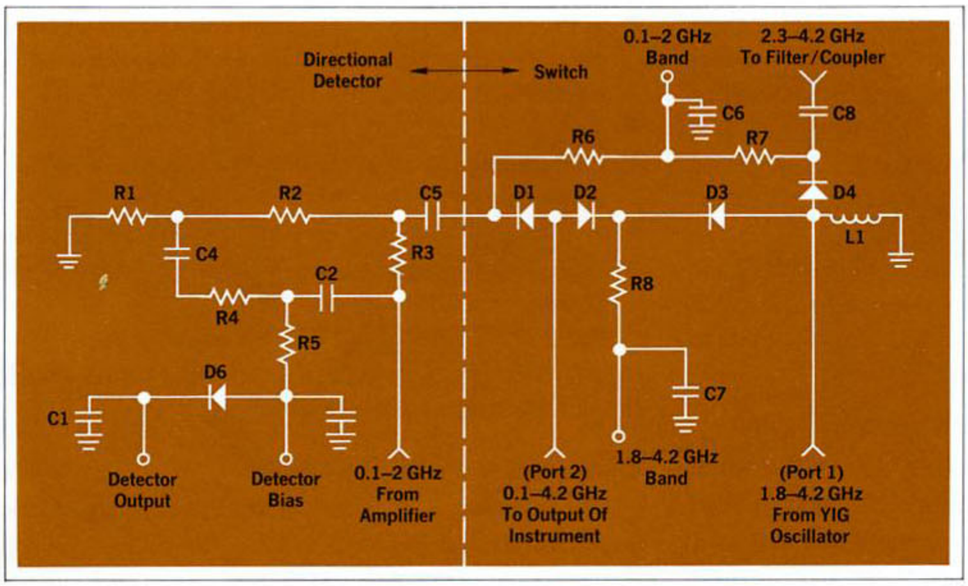


Fig. 2. *Schematic diagram of directional detector/switch microcircuit.*

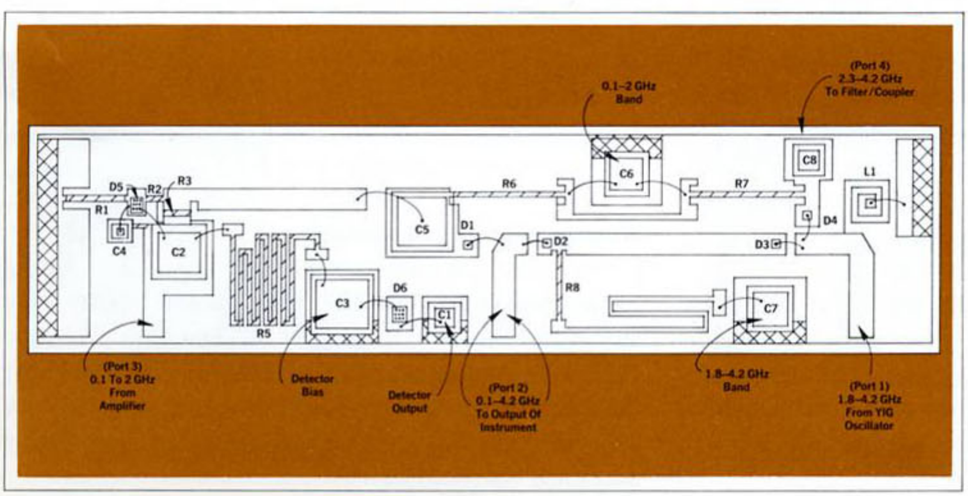


Fig. 3. *Directional detector/switch microcircuit layout.*

Switching is accomplished by diodes D1 through D4. When the 1.8 to 4.2 GHz band is desired, —10 volts is applied to resistor R8. This forward biases diodes D2 and D3 allowing the rf signal to flow from the YIG oscillator to the output of the instrument. When the 0.1 to 2.0 GHz band is selected, diodes D1 and D4 are forward biased through resistors R6 and R7, allowing the YIG oscillator signal to be mixed with the 4.3 GHz local oscillator. The output of the 0.1 to 2.0 GHz amplifier passes through diode D1 to the output of the instrument. Inductor L1 provides a dc return for diodes D3 and D4. The dc return for diodes D1 and D2 is external to the frequency converter package.

During 0.1 to 2.0 GHz operation, isolation between the YIG oscillator and the output of the instrument is an important parameter to minimize leakage of the oscillator to the output of the instrument. Isolation typically greater than 40 dB is accomplished by using 60-mil-square PIN diodes with C_0 less than .05 pF. Fig. 4 shows the insertion loss, VSWR and isolation characteristics between port 1 and 2 of the switch.

The directional detector is diagrammed in Fig. 5.

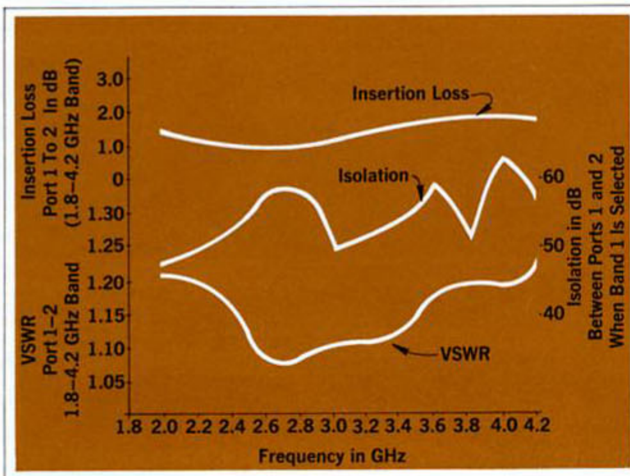


Fig. 4. *Curves describe insertion loss, VSWR, and isolation between ports 1 and 2 of directional detector/switch.*

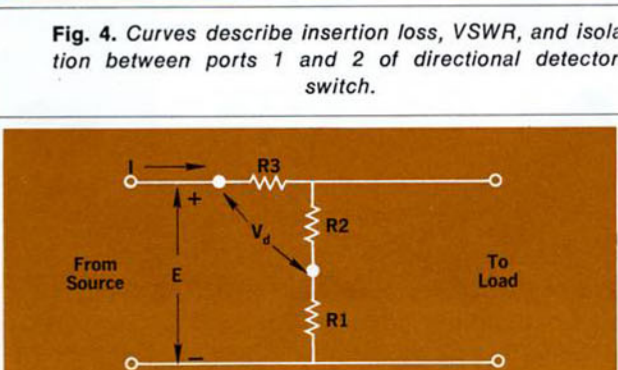


Fig. 5. *Basic directional detector circuit.*

Resistors R1, R2 and R3 form the resistive bridge. Hot carrier diode D5 is used as the detecting diode. Capacitor C4 and resistor R4 are frequency compensating components for hot carrier diode D5, to obtain flat response from 0.1 to 2.0 GHz. Resistor R5 is large to minimize unbalancing of the resistive bridge while allowing diode D5 to be forward biased. The biasing increases the sensitivity of the diode. Diode D6 is identical to diode D5 to temperature-compensate the detected output voltage. Fig. 6 shows a typical plot of the detected output voltage response and directivity as a function of frequency.

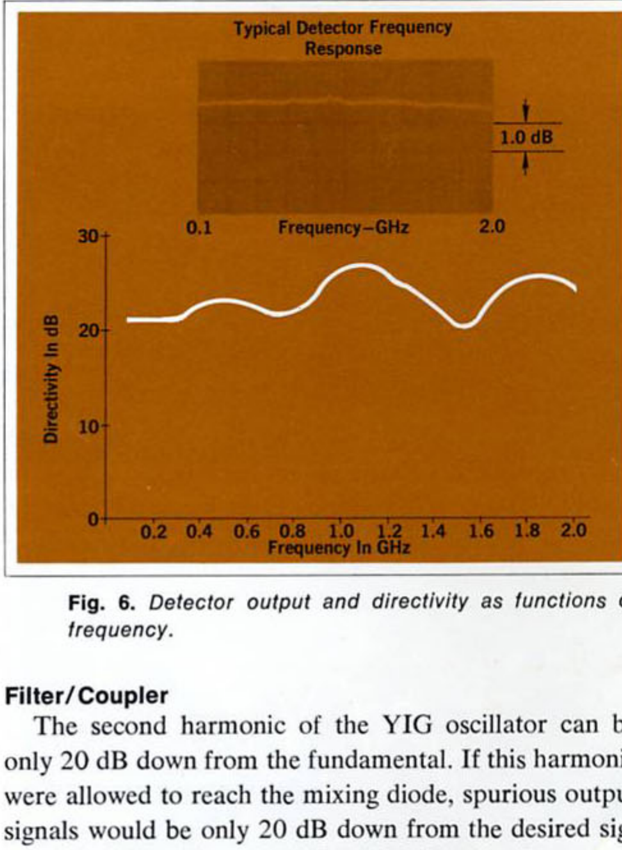


Fig. 6. *Detector output and directivity as functions of frequency.*

Filter/Coupler

The second harmonic of the YIG oscillator can be only 20 dB down from the fundamental. If this harmonic were allowed to reach the mixing diode, spurious output signals would be only 20 dB down from the desired signal. The Filter/Coupler substrate attenuates the second harmonic of the YIG oscillator and combines the low level signal and the 4.3 GHz local oscillator signal for application to the mixing diode.

The 4.2 GHz low-pass filter is a 15-section Chebyshev design. To minimize the inductor capacitance, the ground plane was etched away beneath the inductors.

The coupler provides 10 dB of coupling and approximately 10 dB of directivity. The directivity of the coupler minimizes the interaction of the low level signal with the 4.3 GHz oscillator. Fig. 7 shows the filter/coupler layout and Fig. 8(a) and 8(b) show typical electrical characteristics of the filter coupler.

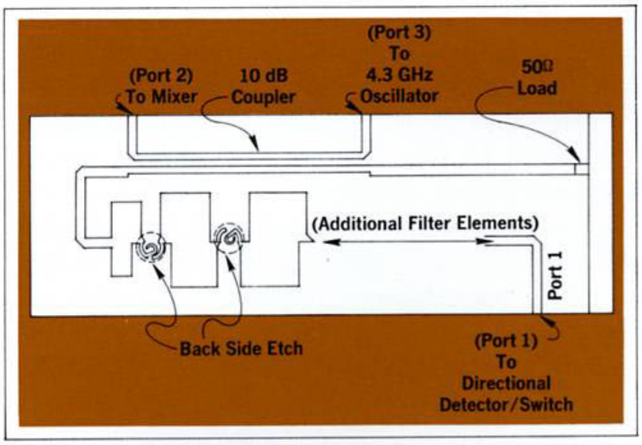


Fig. 7. *Layout of filter/coupler microcircuit.*

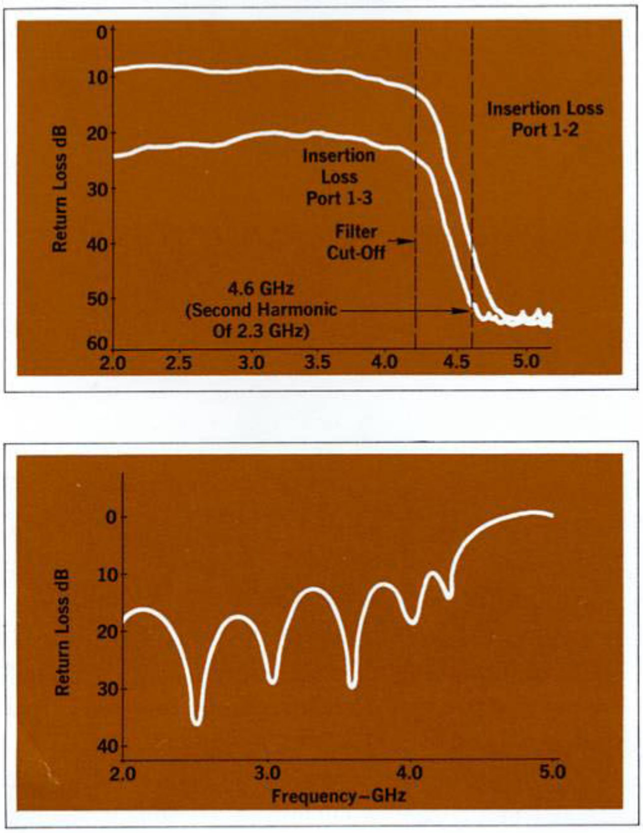


Fig. 8. *a) Insertion loss between port 1-2 and 1-3 of filter/coupler. b) Return loss of port 1 of filter/coupler.*

4.3 GHz Oscillator

The 4.3 GHz oscillator is a push-pull design. This allows the transistors to be operated in the common base configuration; it minimizes the second harmonic output and delivers more rf output power than a single transistor oscillator would provide, using the same type of transistor. A simplified schematic is shown in Fig. 9.

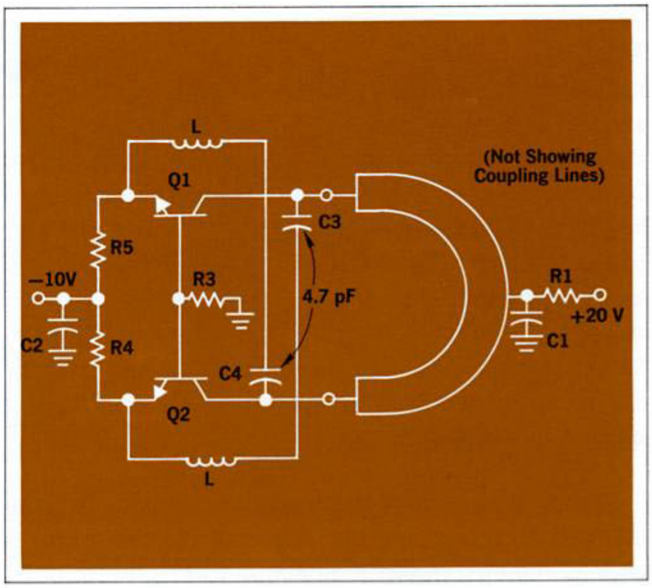


Fig. 9. *Simplified schematic diagram of 4.3 GHz oscillator.*

Analysis is easier if the circuit is transformed into an equivalent unbalanced rf circuit, as in Fig. 10. The push-pull action is replaced here by an ideal transformer which performs the 180° phaseshift.

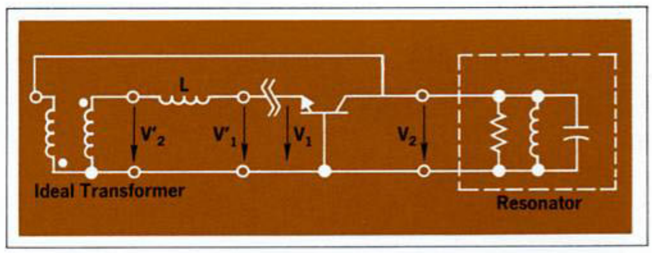


Fig. 10. *Equivalent circuit of 4.3 GHz oscillator.*

The vector diagrams shown in Fig. 11(a) through 11(d) graphically explain why the circuit is able to oscillate. For this purpose the emitter path is broken and a signal V_1 is applied to the emitter. If we make the simplifying assumption that the resonator represents 50 ohms at the resonant frequency, V_2 can be approximated by

$$V_2 = V_1 \times \frac{S_{21}}{1 + S_{11}}$$

where S_{11} and S_{21} = S parameters of the transistor at 4.3 GHz. Push-pull performance transforms V_2 to V'_2. V'_2 sees an inductance L in series with the input impedance of the transistor (represented by a parallel combination of R and L equivalent to S_{11}). This network induces a V'_1 at the starting point of the loop which is in phase with the originating signal V_1 yet is greater than V_1. Therefore oscillation is assured.

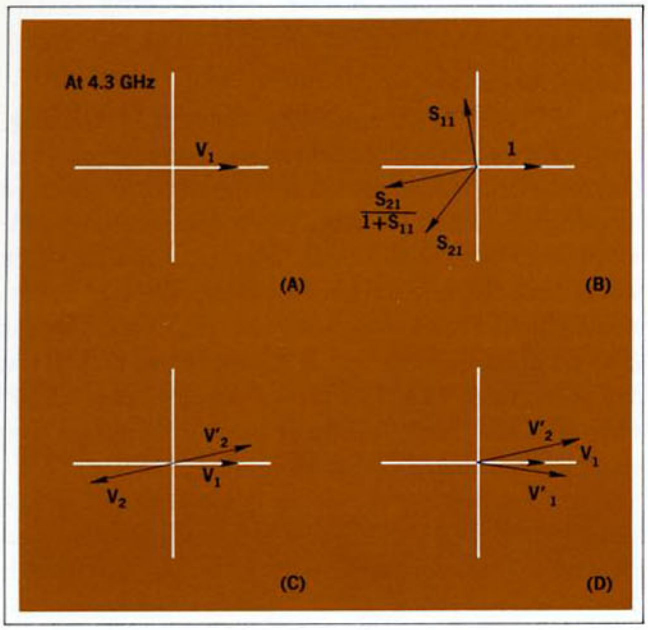


Fig. 11. *Vector diagrams describe behavior of 4.3 GHz oscillator.*

The 4.3 GHz oscillator layout shown in Fig. 12 is built on a ½-inch-square 25-mil sapphire substrate. The two transistor chips are actually mounted on a coupling loop which decouples them from the U-shaped microstrip resonator. The two one-mil wires which connect the ends of the resonator to opposite emitters represent the feedback inductors. The emitter dc potential is blocked from the resonator by two 4.7-pF SiO_2 chip capacitors. To prevent other unbalanced oscillation modes, the common base pad of the two transistors is

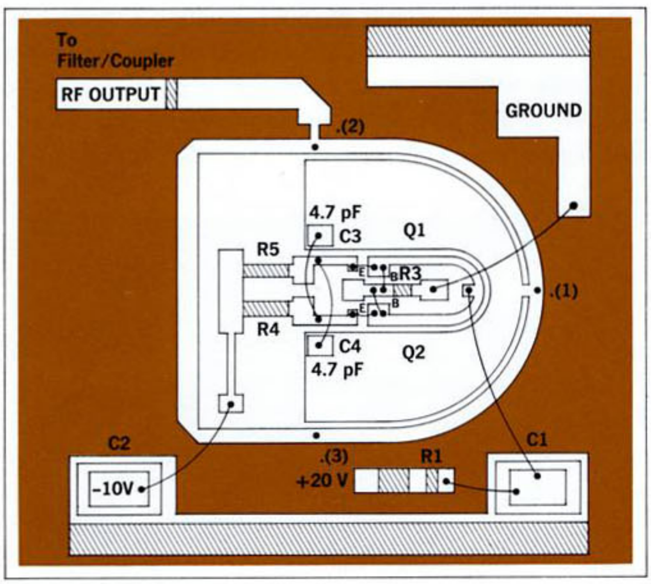


Fig. 12. *Layout of 4.3 GHz push-pull microcircuit oscillator.*

connected to dc ground via resistor R3. Thus it acts only as ground when the devices are working push-pull.

The resonator is surrounded by a microstrip-line which couples out the rf. Point 1 is the floating ground of the resonator with zero rf voltage and maximum rf current. Maximum power is obtained if the short path from 1 to 2 is $\lambda/4 \times (2k + 1)$ and the long path from 1 to 2 via 3 is $\lambda/4 \times (2k + 3)$ ($k = 1, 2, 3$----, $\lambda =$ wavelength on sapphire). This unbalances the symmetrical signals and adds them at point 2.

The output power obtained is typically +10 to +13 dBm. Frequency drifts at a rate of —500 kHz/°C. This is caused mainly by the high temperature drift in the dielectric constant of the sapphire. Harmonics are typically more than 40 dB below the fundamental.

Fine tuning of the oscillator is accomplished by varying the collector voltage. Changing collector voltage varies the collector-to-base capacitance which in turn affects the frequency of oscillation. The —500 kHz/°C frequency shift of the oscillator is compensated by providing a collector voltage supply which increases linearly with increasing temperature.

Mixer

It was possible to meet the spurious objective (30 dB down from the fundamental) with a single hot carrier diode. This eliminated the need for a more complex and expensive balanced mixer configuration. The mixer diode is placed on a separate 0.5 x 0.125 x 0.025 inch high alumina substrate. Since no resistors were required on this substrate, high alumina material could be used in place of sapphire. This reduced the cost of the mixer. Placing the mixer diode on a separate substrate also allows for easy replacement in case of diode failure.

Fig. 13 shows the layout of the mixer. Tuning stub (1) grounds one side of the diode while acting as a tuning element for the mixer.

2 GHz Low Pass Filter

The 2 GHz low pass filter attenuates the higher order mixing products and the low level and local oscillator signals. The filter, constructed on a 50 mil sapphire substrate, is of elliptical design to achieve extremely sharp cut-off characteristics since the filter must pass 2.0 GHz, yet highly attenuate higher frequency signals starting at 2.3 GHz. Fig. 14 shows the filter layout and Fig. 15 shows typical filter cut-off characteristics.

The YIG-tuned transistor oscillator[1] for the 8633A Oscillator Module and the wideband amplifier[2] for the 8633A Heterodyne Module have been described elsewhere.

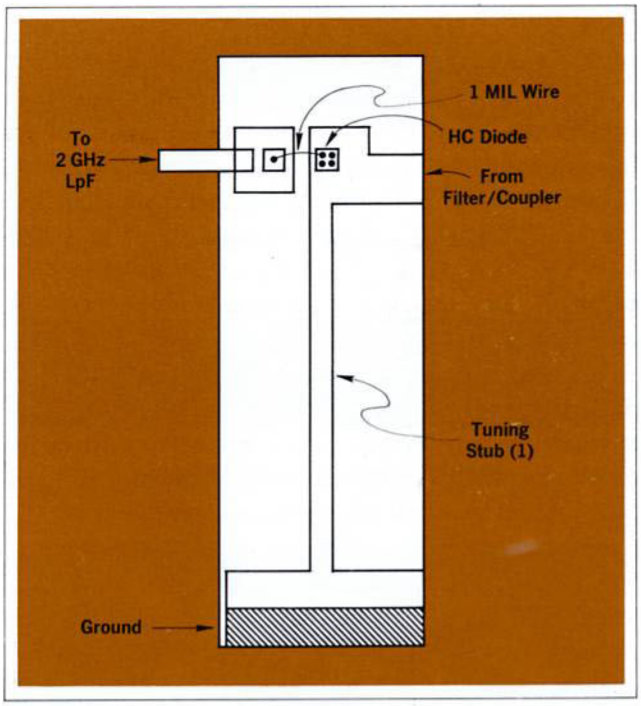


Fig. 13. *Layout of mixer microcircuit.*

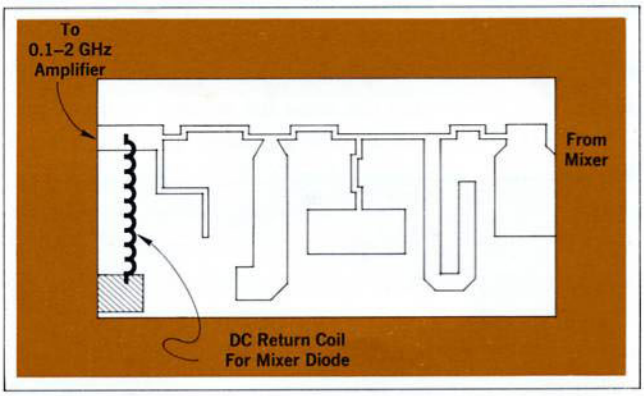


Fig. 14. *Layout of 2-GHz low pass filter microcircuit.*

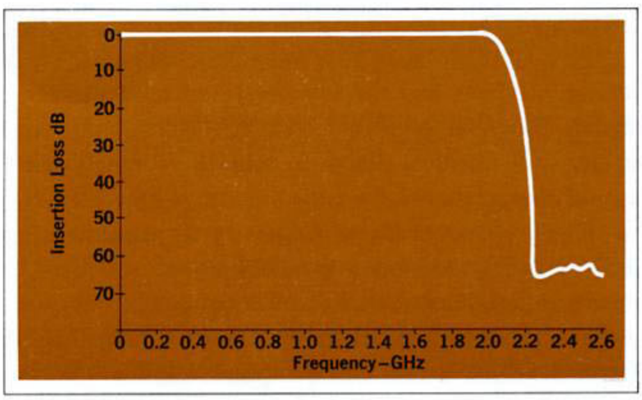


Fig. 15. *Frequency response of typical 2-GHz low-pass filter.*

Microcircuits for the 1.8-to-4.2 GHz Oscillator Module

The PIN Modulator

A new PIN diode modulator provides rf blanking, amplitude modulation, and automatic level control. Thin-film microcircuit construction and the use of unpackaged diodes having small parasitic reactances made it possible to produce a lumped circuit attenuator design which works from 0.8 to 4.2 GHz. PIN diodes are used as current-controlled microwave resistors. The new circuit has a 60 dB dynamic range with a VSWR of less than 2:1 at any attenuation, and is built on a sapphire chip 0.230″ wide by 0.460″ long; it is many times smaller than conventional distributed-circuit modulators.

Fig. 16 shows that the circuit is a bisected Pi attenuator formed by diodes D1, D2, D6, and D7 which provide 40 dB of attenuation. D3, D4 and D5 provide attenuation from 40 to 60 dB. For small attenuation, D2 and D6 are forward biased by driving a negative current into the control lead. For maximum attenuation, a positive control current is applied, thus biasing D3, D4, and D5. Any reflected signal from the shunt diodes is attenuated twice by D2 and D6 which are essentially at zero bias due to D1 and D7 drawing current from the plus supply. At this point D1 and D7 are biased to have a 25 ohm rf resistance which together with the 25 ohm fixed resistors give a 50 ohm input and output impedance. For the first 40 dB of attenuation, the rf resistances of D1, D2, D6, and D7 closely follow the optimum values for a Pi attenuator thus producing a well matched device. The smaller RC networks in shunt with D2 and D6 minimize the effects of diode junction capacitance.

Before the first breadboard version of this modulator was built, various circuit elements were characterized separately, and an equivalent circuit was analyzed on a time-shared computer. This analysis made it possible to

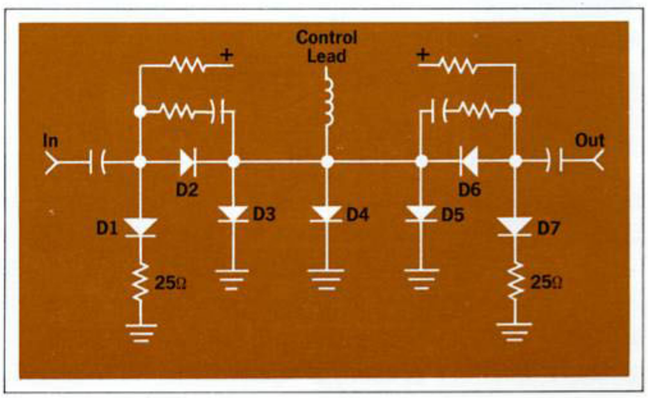


Fig. 16. *Equivalent circuit of PIN diode modulator-attenuator.*

adjust circuit parameters to achieve the final performance. The decision to use three shunt diodes was made when the equivalent circuit would not produce sufficient attenuation with fewer diodes. The effect of the RC networks was also predicted by the computer model. A special non-linear analysis program was used for optimizing the values of the positive dc supply and the bias resistors, to provide the best possible transfer function and impedance match. Fig. 17 shows some of the equivalent circuit elements used for the computer analysis. It is interesting to note that even a simple thin-film resistor must be analyzed as a transmission line to perform accurately in the circuit model. The values shown were either directly measured or verified experimentally.

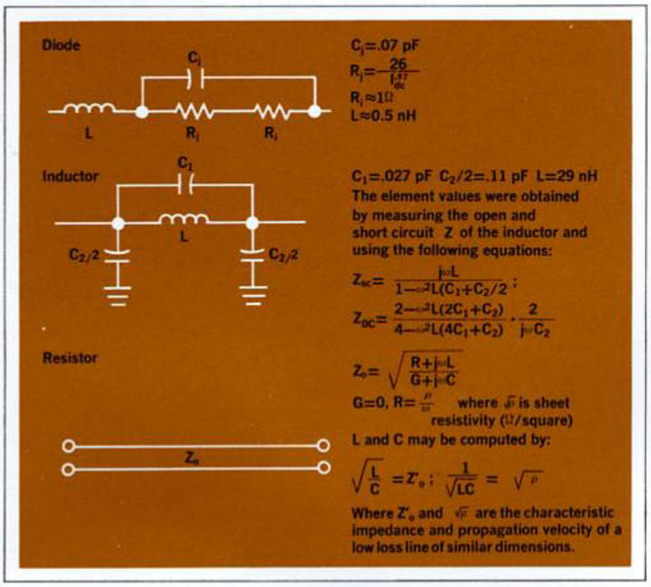


Fig. 17. *Equivalent circuit elements of PIN modulator-attenuator.*

The use of series and shunt diodes shows a unique advantage when the overall ALC loop is considered. If the transfer gain of any modulator is defined as the change in control current necessary to provide a certain change in rf power, it can be shown that the gain of most shunt-diode modulators varies approximately directly with the amount of attenuation introduced. For example, the current change necessary to vary the rf power from —20 to —22 dB will be about 10 times larger than it was going from —10 to —12 dB. Reducing loop gain would degrade ALC performance as the leveled power output is reduced by changing the rf output control, thus causing more power variations and poorer source match. The series shunt modulator avoids this problem by obtaining most of its dynamic range with very low bias currents. Typically there is about 10 dB change over the first 30 dB of dynamic range, thus providing effective leveling over the entire range of the power level adjustment.

1.8–4.2 GHz Directional Detector

The basic function of a directional detector is to sense rf energy from one direction without regard to other signals, and produce a dc voltage proportional to the rf amplitude. The device can be broken into two sections for analysis: a directional coupler plus a crystal detector. For wideband applications, these two units may be developed separately, then combined to give the desired performance. For this microcircuit unit it was necessary to compensate the directional coupler to achieve the desired output from the combined coupler/detector unit.

The 'directional' qualities of the device are completely a function of the coupler directivity. As can be seen in Fig. 18, it is desired to sense a signal at port 3 for a wave incident at port 1, but to receive no signal at port 4. Conversely, a signal incident on port 2 would not transmit energy to port 3. In actual units, there is a very small error signal at the 'isolated' port, and it is convenient to define the coupler directivity as 10 log $\frac{P_1}{P_3}$ for the same conditions. It is evident, then, that the directional coupler can be used to sense a forward wave while ignoring (or almost so) the reverse wave (the wave incident on port 2 as the result of a reflective load, etc.).

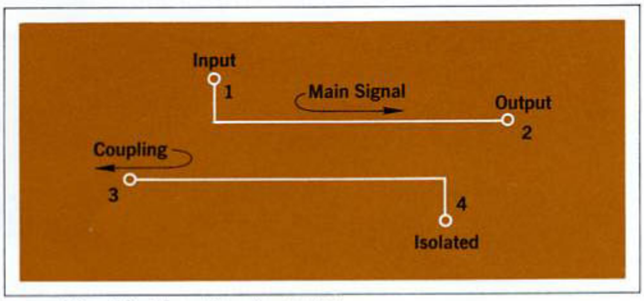


Fig. 18. *Directional coupler.*

The second section of a directional detector is the crystal detector unit (diode detector). This unit senses rf energy and converts it to a filtered dc signal. A second tracking diode is added for temperature compensation of the unit. The diodes are Schottky barrier types and require a dc bias to lower the junction resistance for rf to dc conversion. Fig. 19 shows a schematic.

When the two sections are combined, it is necessary to adjust coupling characteristics to first compensate for an inherent rise in dc output with increasing frequency of the detector circuit, and secondly to compensate for cable losses before the output signal reaches the front panel. This gives the unit a —.75 dB frequency response from 1.8–4.2 GHz with a minimum directivity of 26 dB. The

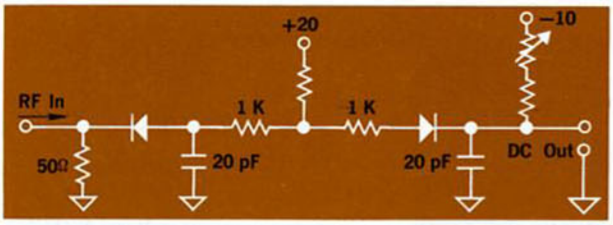


Fig. 19. *Temperature-compensated crystal detector.*

coupling has a maximum ripple of ±.25 dB.

Inside the directional detector one finds some new ideas for thin film integrated circuits. Because a directional coupler is a precision device, simple microstrip geometrics will not meet all the electrical requirements. First, ground currents that are interrupted tend to launch higher order modes inside the package, which leads to poor directivity. Secondly, mixed dielectric media had two large error signals due to mismatched mode impedances. Again the directivity is adversely affected. To solve these problems, a symmetrical structure called 'triplate' was incorporated in a unique sandwich-structured box (Fig. 20) using two dielectric substrates instead of the usual single unit.

It should also be mentioned that high alumina substrate material was used rather than synthetic sapphire

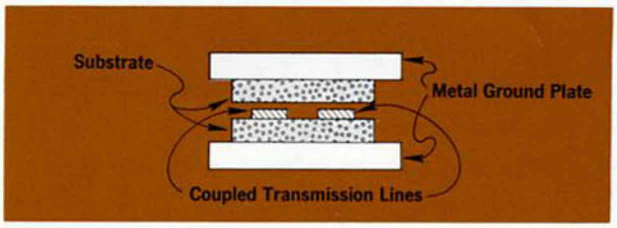


Fig. 20. *'Triplate' sandwich structure used in microcircuit directional detector.*

because of the anisotropic properties of sapphire (the magnitude of the dielectric constant is a function of the direction of energy propagation in an anisotropic medium).

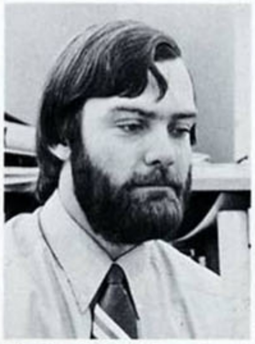
Acknowledgment

Much credit is due to Cyril Yansouni for design concepts for the 8632A Heterodyne Module. Dieter Scherer was responsible for the original design concepts for the 4.3 GHz local oscillator used in the frequency converter package. Special thanks go to Bob Jacobsen for the design of the frequency converter package.

Ronald E. Pratt

Ron Pratt received his BSEE in 1967 from the Newark College of Engineering the hard way—at night school while holding a full time job at Bell Telephone Labs. Ron was responsible for development of the PIN Modulator for the 8633A Heterodyne Oscillator. Night school still fills a large part of Ron's time with teaching of Microwave Electronics two nights a week at Foothill College. High fidelity and a gleaming Porsche for autocross racing help complete Ron's busy schedule. Ron belongs to the Microwave Theory and Techniques Group of IEEE.

Robert W. Austin



Soon after graduating in 1968 from Seattle University with a BSEE, Bob Austin joined Hewlett-Packard. He immediately joined the team designing the 8620 Sweep Oscillator and took sole responsibility for the directional coupler. Bob is now involved in packaging design and microcircuit processing at HP.

This fall, Bob began studies toward an MSEE at Stanford. Racing his TR4 sportscar is Bob's idea of an exciting weekend. So far, he has only been in autocross races but he plans to road race next year. Bob is a member of IEEE's Microwave Theory and Techniques Group.

Arlen E. Dethlefsen

Arlen Dethlefsen joined HP in late 1968 after five years with Bell Telephone Labs where he helped develop microwave radio systems. Arlen received his BSEE and BS in mathematics from California Polytech in 1961 and an MSEE in 1963 from Northwestern University. As project supervisor, Arlen was responsible for microcircuit development for the 8632A Heterodyne Converter. Skiing takes up much of Arlen's time in the winter months while his summer pastime is playing tennis.

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