



microwave JOURNAL

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The 1982 MTT-S Symposium
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ON THE COVER: The Hyatt Regency, headquarters of the 1982 IEEE MTT-S Symposium/Exhibition, dominates the foreground of a view of the Dallas skyline. Cover photo: R.O. Bumpass. Also featured is Wiltron's new 40 mW 2.0 - 18.0 GHz sweeper and its Cover Story begins on page 164.

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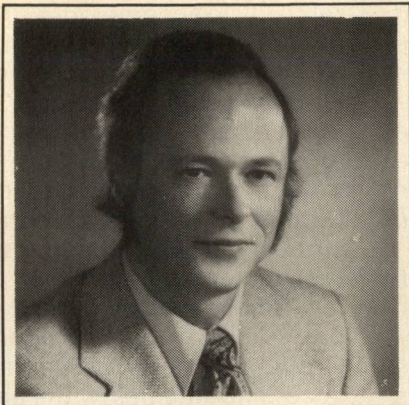
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"No duplexer is needed. We turn on the 450 kW average power transmitter for 20 minutes, mechanically switch over, wait 2 hours for the echo, then receive for 20 minutes." So did Dr. Donald Campbell, Acting Director of Cornell University's Arceibo Observatory in Puerto Rico describe the facility's use to illuminate and receive radar echos from Saturn's rings, a *one billion mile range* radar target! What prompted the development of a radar with a 1000 foot diameter (20 acre) antenna (Figure 1)? What logistics influenced its design and construction? How is it used today and what can it tell us?

Mapping the Ionosphere

In the early 60's understanding ionospheric radio properties was critical to military communications. This required mapping the plasma density of the ionosphere, which extends roughly 30-300 km above the earth's surface. Prior to that time such mapping could be performed only by transmitting vertically directed signals in the 2-15 MHz range and listening for the plasma density determined resonant reflection. High reflection occurs in the ionospheric layer at which the plasma resonance coincides with the probing frequency. In this manner the plasma elevation profile could be plotted from the round trip path time of the probing pulses. How-

ever, the plasma density peaks at various altitudes, depending upon season, temperature, and related conditions. For elevations above these peaks, the density could not be so mapped because echos from the corresponding densities below the peaks would mask the returns from above.

A high peak power highly directive radar was needed using frequencies high enough that the beam suffers negligible attenuation or refraction as it passes through the entire ionosphere; backscatter from the free electrons as a function of time then maps the ionosphere's plasma density^{1,2}. Figure 2 shows a typical mapping.

Currently the Arecibo radar can

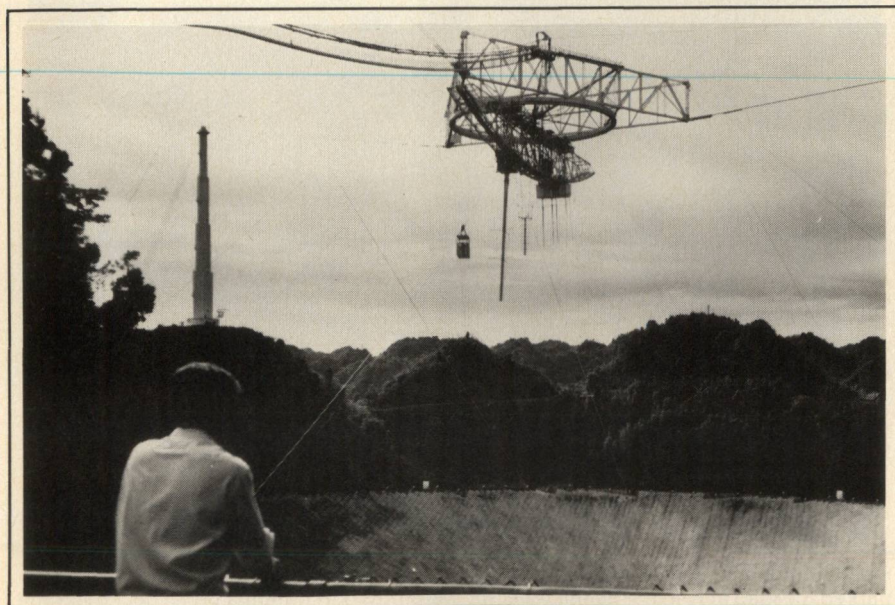


Fig. 1 Dr. Campbell watches as personnel gondola (center) carries technicians and scientists to 600 ton feed assembly supported 450 feet above spherical dish antenna.

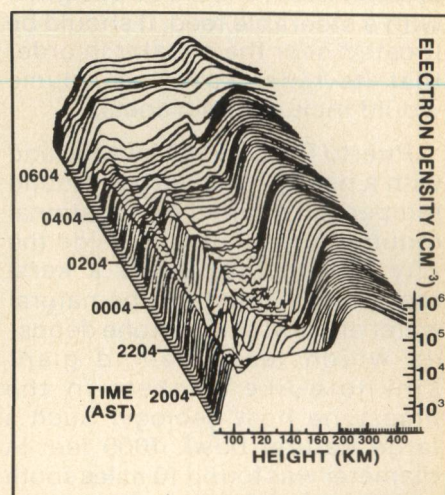


Fig. 2 Electron density map at 15 minute intervals made of the ionosphere using the Arecibo radar². Sunlight causes the density increase after 0600. Low altitude variations are due to winds and electrical fields.

be operated from 50 to 5000 MHz, well above the plasma resonant frequencies and therefore independent of them. The Advanced Research Projects Agency (ARPA) funded Cornell University to design and build this ionospheric sounding radar. What resulted was an instrument capable of probing deeper into intergalactic space than any instrument built before or since then; but this spinoff was the result of a courageous system gamble, a bold stroke requiring departure from conventional antenna design to use something which had never been employed in a large radar antenna before, a *spherical* shape. We'll see more about this a little later, but first consider the requirements of this unique radar application.

Site Location & Development

Even using the most powerful transmitters available, 2.5 MW peak, 150 kW average at 430 MHz, a large antenna aperture was needed. Such an aperture would be too large to fabricate economically on level ground. Rather a depression in the earth would be used, conforming to the desired antenna shape. To achieve adequate sensitivity, acres of antenna area were needed, so a natural depression, preferably with three surrounding high hills from which a feed structure could be supported was sought. If such an antenna dish, too large to be steered itself, should be designed with a steerable feed, it should be located near the equator in order that its radar scanning volume would include the planets.

Puerto Rico, at 18° latitude and with a nearly constant year round temperature suited the locational requirements. There, outside the city of Arecibo, exists a karst region, so named for the natural water erosion of limestone deposits which leave behind giant sinkhole-like pockets in the limestone host geology. Such a large natural bowl, 1000 feet in diameter was found 10 miles south of the city of Arecibo. Two of the desired three hills were present, the third support would come from a 350 foot concrete tower. The natural depression's shape con-

formed, within 20 feet, to a perfectly spherical surface with a 900 foot radius of curvature. Located in a low mountainous range, the temperature is always within 8° of 80° Fahrenheit. Below the depression, limestone caves were already situated to drain the canyon of the torrential Caribbean rains. The perfect site had been located.

While perhaps ideal from the antenna engineer's perspective, the site posed formidable construction problems. Small two lane roads were the only access in this underdeveloped area. Miguel Feyjoo, head of the electronics department and Domingo Albino in charge of transmitters, recall that 10,000 cubic yards of concrete (weighing over 2000 tons) and 1000 tons of steel would be consumed before the project was completed. Six hundred tons of the steel would comprise the feed platform having to be supported 480 feet in the air above the bowl on three cables. The cables, in turn, would be secured to a three tower tripod, the longest leg of which would exceed a height of 35 stories. Crews of up to 250 people labored for two years, this at a time in the early sixties when wages were only \$1.10 an hour. The \$9M completion price, a bargain even then, represents over 500 man-years' of effort.

The antenna formation was begun by rounding out the natural cavity to accommodate the 1000 foot diameter reflecting surface. Although the depression was already nearly spherical, "truing up its shape" involved slicing 10 to 20 foot thick limestone "skins" from the surrounding hill faces (Figure 3), often to a height of 12 stories. As with construction of the ancient pyramids of Egypt, few details of this project would prove to be trivial assignments. Drainage, as noted earlier, is accommodate below the antenna surface by natural caves, but the heaviest rains, for which the tropics are well known, require assistance from a 200 horsepower auxiliary diesel pump to keep the water level from flooding the facility.

Surrounding the 1000 foot aperture a concrete rim was poured and fitted with one inch thick steel, suspension cables. In a crisscross weave of 3 by 25 feet these are attached to the rim with massive turnbuckles. Initially mesh screening was placed on the cable weave and the resultant "dish" surface provided an antenna operable at 430 MHz. Later, however, the entire surface was upgraded with aluminum mesh panels extending its frequency operation to C-band. Twenty acres of panels, over 38000 of them,

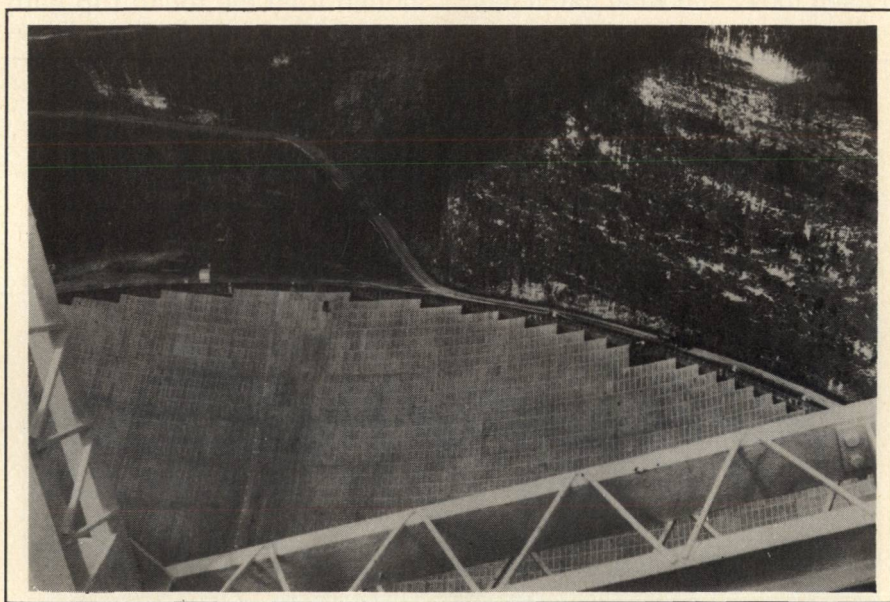


Fig. 3 Shaped limestone hills 12 stories high surround the antenna. Jeep with trailer is parked on road at rim of antenna reflector. Missing 3' x 7' antenna panel near outer rim provides larger than a man sized perspective.

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each measuring 3 x 7 feet were needed. For the higher frequency operation a better calibrated surface also was required, this necessitating that a crew of surveyors, 6 to 8 at a time, toil for over a year, twisting the 2000 turnbuckles and 38000 adjusting studs (Figure 4) of the giant trampoline-like antenna surface until the final surface perfection, true to within 0.12 inch RMS finish, was achieved

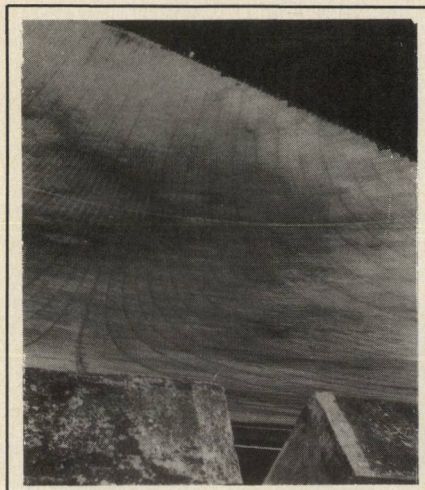


Fig. 4a Antenna surface of perforated aluminum panels is supported by 1" diameter steel cables.

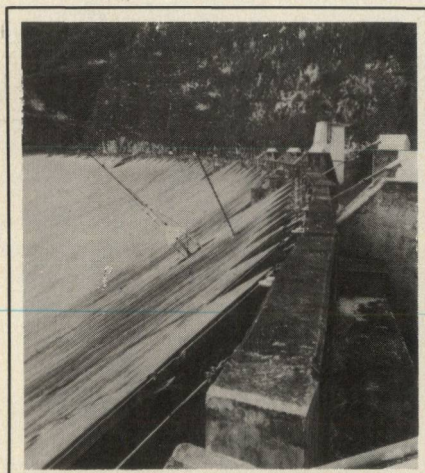


Fig. 4b Heavy turnbuckles and concrete rim support 20 acre reflector.

throughout the three quarters' of a million square feet of area that the 1000 foot diameter surface provides.

The Spheric Antenna Decision

Antenna designers, led by Cornell's Dr. William Gordon, were faced with a most difficult decision. The task at hand, for which the ARPA funding was made available, required only a radar that would operate for a target

directly overhead, namely the ionosphere. There was no need for steering. However, such a giant unsteerable instrument, its ionospheric studies completed, would no longer be needed. The vast labor investment would have been expended for a one-shot experiment. Rapid obsolescence would be the fate of a conventional parabolic reflector, which focuses all of its energy to a single point, a reflector geometry so greatly simplifying the antenna feed system that is is virtually the only kind of antenna surface used for large radar antennas even today. Such a paraboloid antenna surface would be too large to steer physically, and it cannot be scanned by moving its feed.

A spherical surface, on the other hand, would permit scanning by steering the feed. As shown in Figure 5, plane waves incident on the spheric surface are focused along a line parallel to the direction of their origin.

The catch, however, is that the spheric focuses its energy not to a point as does the parabolic, but rather along a line which parallels the incident rays. Therein was the challenge. Even for targets directly overhead, an antenna feed would be needed which could collect and properly phase energy focused along a line that, for an antenna of this diameter, would be 95 feet long to collect 70 percent of the aperture's reflected power! Along this "focal line" energy from various annular rings within the spheric antenna aperture converge at corresponding focal "points" along that line.

Worse yet, all of the energy arriving from one of these annular sections does *not arrive with the same polarization*. To visualize this, consider that the entire spheric aperture is illuminated by a linearly polarized electromagnetic plane wave arriving from space. Energy which arrives at any given point along the sphere's focal line comes from a corresponding annular zone in the antenna and contains electric field polarizations which assume all linear orientations, about a 360° rotation at the line feed collector as shown in Figure 6. In this illustration a slotted waveguide col-

lector ("feed") has been assumed as the means used to gather the line-focused energy.

To combine the energy, coupling holes are needed at the top, sides and bottom of the guide. As the linearly polarized E field impinges on the entire array aperture, portions of that energy which reflect from the top of the antenna to a part of the line feed will strike the slotted waveguide feed so as to enter via top-wall (broadwall) coupling holes (Figure 6).

But energy reflected from the sides of the antenna will enter through side (sidewall) holes, and energy arriving from the bottom of the spheric surface through a bottom broadwall hole. The prob-

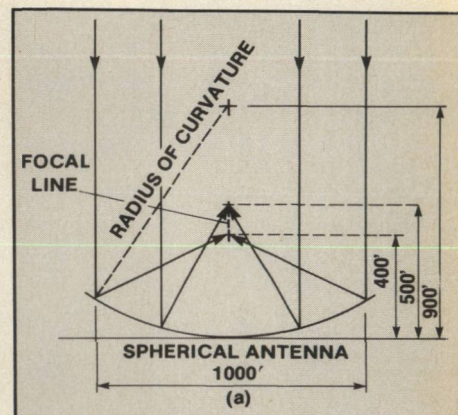


Fig. 5a 10000 foot diameter antenna with 900 foot radius of curvature for which 70% of area focuses to approximately 100 foot line 450 feet above base.

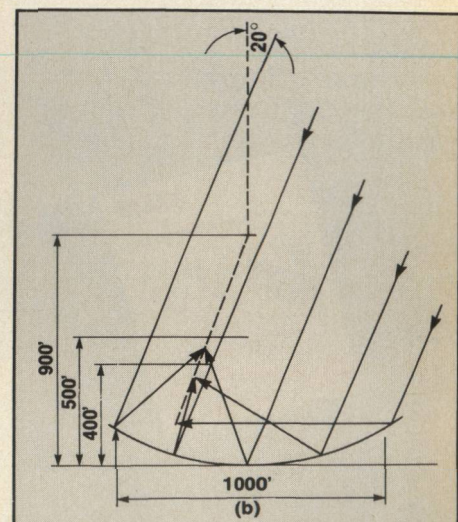


Fig. 5b Spheric antenna scanned 20 degrees off zenith (boresight) direction focuses along a line parallel to scanning when reflecting "dish" is stationary.

[Continued on page 116]

Many Frequencies and Components

The Arecibo Observatory facility can be used as a radar with transmission and reception cities at both 430 and 2380 MHz. The CW S-band transmitter consists of a Varian klystron with 2.5 MW peak, 450 kW average power capability. A dual channel receiver uses a maser amplifier. This consists of a three level ruby crystal (Al O) with 0.05% chromium (+3) ions and cooled to a liquid helium temperature, about 4.5° Kelvin. This amplifier uses a 12.9 GHz pump power source at 100 mW.

The UHF radar transmitter, employing two klystrons for 2.5 MW peak, 150 kW average power, is located in the ground station adjacent to the antenna, and the microwave energy plumbed to the feed structure via a waveguide line. However, at S-band the losses of such a transmission system would be prohibitive, therefore the entire transmitter is located in the cabin structure on the feed

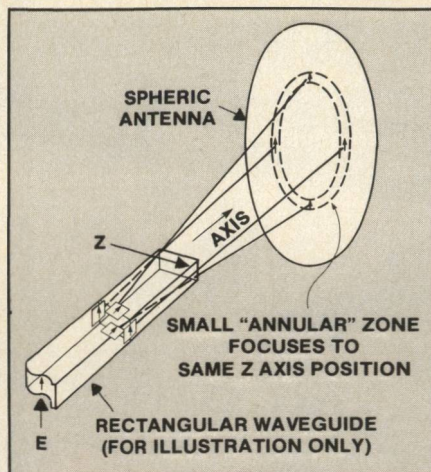


Fig. 6 Illustrating polarization rotation at waveguide feed when antenna is illuminated by a vertically polarized planar E field.

lem is that *broadwall coupling holes produce radiated signals which are 90° out of phase compared with sidewall slots!*

Of course, for symmetry reasons, circular rather than rectangular cross section waveguide would be used, but the phase shift just described relative to the vertically directed resultant travelling field in the waveguide is the same.

Accordingly, in order for the energy to combine perfectly within the waveguide, these signals must be brought into phase, utilizing some form of spatial phase shift which varies continuously according to angular orientation about the feed. Such antenna problems and their solutions had never been treated previously on a scale such as this. Moreover, research for the development of the feed would have to be done in parallel with the antenna's construction!

Nonetheless, the decision was made to proceed with an antenna whose pointing could be steered via the feed. A breakthrough, occasioned from the early work of Gustincic and Love³ consists of dielectric rings (Figure 7) placed in the vicinity of the waveguide feed's coupling holes.

The rings serve as waveguides below cutoff for one E field orientation (polarization) and above cutoff for the other. With proper design the required 90° phase shift is obtained as a function of the angular position about the line feed collector.

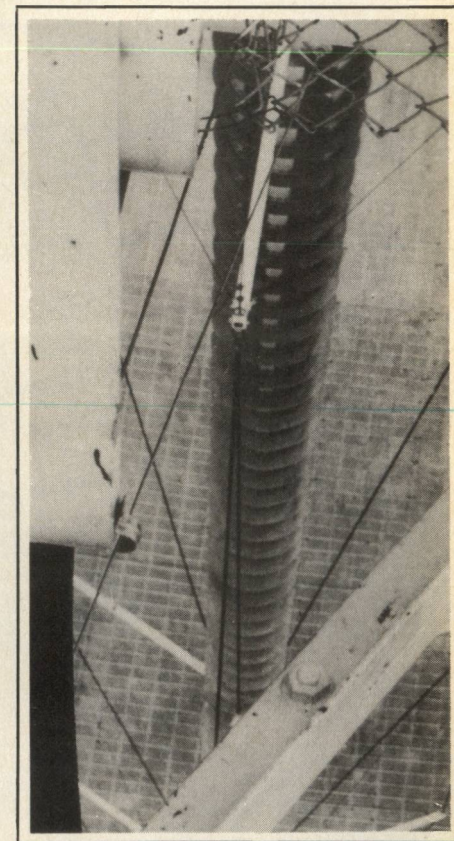
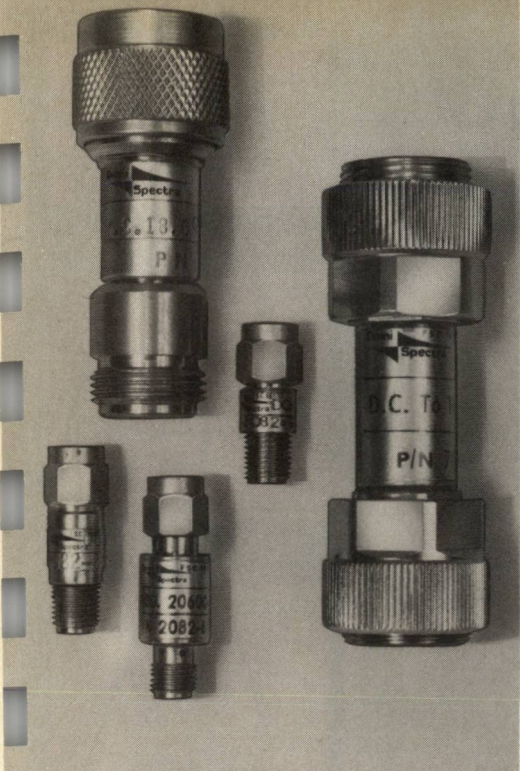


Fig. 7 The 100 foot long feed uses dielectric compensation rings to co-phase the variable polarization reflections from the dish.

[Continued on page 118]



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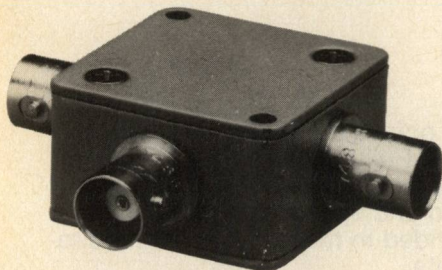
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[From page 116] RADAR

platform. Figures 8 and 9 show the high power S-band transmitter tube and its 450 kW average power water cooled load.

These components are located in the carriage house which moves along the bow shaped feedarm to orient the feed along the line parallel to the direction to which the antenna is to be scanned. The second degree of motion, consisting of rotation, is effected by the round track which permits a cylindrical coordinate rotation of the entire transmitter cabin and feed assembly (Figure 10). The combined motions permit aligning the line feed located below the transmitter cabin along any direction within a cone whose apex is plus or minus 20° about the zenith direction for the antenna.

One of the most important microwave frequencies for a radio telescope is at 1420 MHz, the hydrogen line frequency. Previously, a receiver designed for this frequency range to achieve low noise figure would be built as a parametric amplifier. Presently, a cooled GaAs FET amplifier provides a 40° Kelvin system noise temperature (noise figure of the receiver is about 15° K) and much more reliable and practical operation. Figure 10 shows the rack mounted dual channel amplifier and its liquid helium refrigerated system.

Details of the design of this amplifier utilizing a special quartz bulb transition into waveguide to minimize both ohmic and thermal losses are given in an accompa-

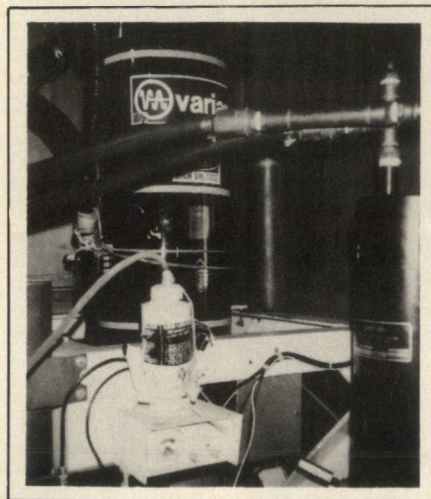


Fig. 8 S-band klystron delivers 2.5 mW peak power.



Fig. 9 Water cooled S-band load in feed cabin can absorb up to 450 kW of average power. Cooling radiators can be seen on the side of feed in Figure 10.

nying paper by George Peter in this issue.

Next The Universe

Just how far away can a signal be detected by the Arecibo radio telescope? To answer that question, radio astronomer Ken Turner responds simply that "that varies according to what you are listening to." Consider the 1420 MHz hydrogen line frequency. Radio emission at this frequency occurs in a neutral atom of hydrogen whenever the electron flips its spin orientation from parallel to antiparallel to the spin of the proton in the atom's nucleus. The Milky Way galaxy, in which we live, behaves as though it contains a uniform gas of such spin flipping electrons with a resultant peak radio noise temperature of 100° Kelvin. Rotation of our galaxy at 300 km per second produces a corresponding doppler shift in the 1420 hydrogen line frequency of about plus or minus 1 MHz. Accordingly, a receiver bandwidth of about 2 MHz centered at 1420 MHz would be needed to listen for a galaxy similar to our own.

The Arecibo antenna beamwidth is about 0.05° at 1420 MHz and the solid angle of this beam is therefore about 0.002 square degrees. As we shall see, the Arecibo receiver easily can detect a signal which is 0.1° Kelvin above the background noise of outer space. This means, therefore, that we could see a galaxy similar to our own in size and hydrogen gas density when it subtends about 1/100th of the beam of the antenna, or 1/2 minute of arc. Our galaxy is 100,000 light years across. The

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[From page 118] **RADAR**

tangent of 0.5 minutes of arc is equal to $1/6875$. Accordingly, the Arecibo instrument can easily detect a galaxy like the Milky Way at a distance of 700 million light years!

The RMS noise for such an observation is given by the noise temperature of the receiver divided by the square root of the product of receiver bandwidth and integration time. With the 40° Kelvin noise figure, 200 KHz of bandwidth, and a five second integration time, the inherent noise of the system would be 0.005° degrees Kelvin. From this it is apparent that the one tenth degree Kelvin assumed signal level from the distant galaxy would be discernable readily, being 20 times stronger than the background noise.

Distances are not noted in miles by astronomers because the distances that they encounter are so large. An important unit of distance to astronomers is the "parsec", this being the distance at which the radius of our orbit around the sun (93 million miles) would subtend an angle of one second of arc. One can see that this distance measure has historical practical significance, since observation of an object at 1 parsec's distance for two readings spaced 6 months apart would result in an angular telescope

motion of 2 arc seconds, about the smallest practical resolvable angular celestial difference. It turns out that one parsec is equal to about 3.2 light years; and, therefore, our hypothetical galaxy example having 0.1° Kelvin signal strength would be 220 megaparsecs away.

Most astronomers believe in the "big bang" theory by which it is theorized that all objects in the universe are diverging from some central explosion. Measurements of the optical doppler shift (called the "red shift" because objects moving away from us have their color spectra shifted towards the longer, red wavelength) follow the Hubble Constant Law, in that they have (and have had since the universe's big bang origin) a velocity relative to ours of 50 km per second for each megaparsec of distance. Accordingly, the 0.1° Kelvin signal from a galaxy which we would observe at 220 megaparsecs (and galaxies are frequently observed at such distances) would have been moving away from us since the beginning of time at a relative speed of 7 thousand miles per second, 4% of the speed of light.

Of course such red shift distance measurements require that the object be observable optically. In fact, some unusual galaxies are observable with greater sensitivity by radio than by optical telescopes. Quasi stellar objects (quasars) are observable at what are believed to be much farther distances.

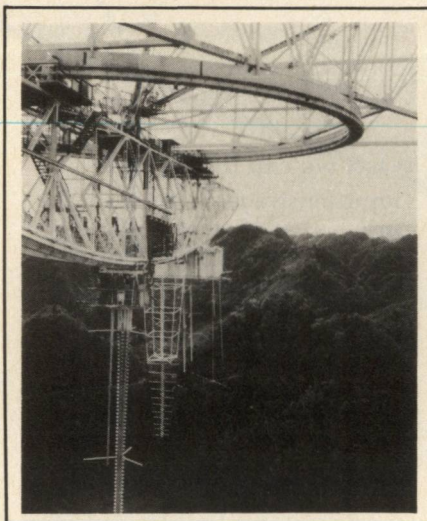


Fig. 10 Steering mount for spherical antenna has 360° rotation and $\pm 20^\circ$ axial motion as feed moves along bowshaped feedarm. White cabin houses S-band transmitter. Note staircase left of center leading to catwalk at upper right for size comparison of 600 ton suspended feed platform.

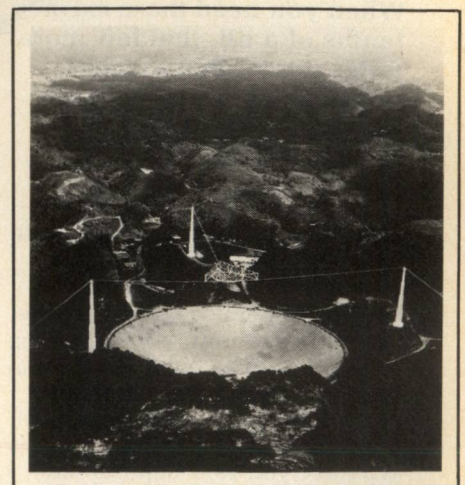
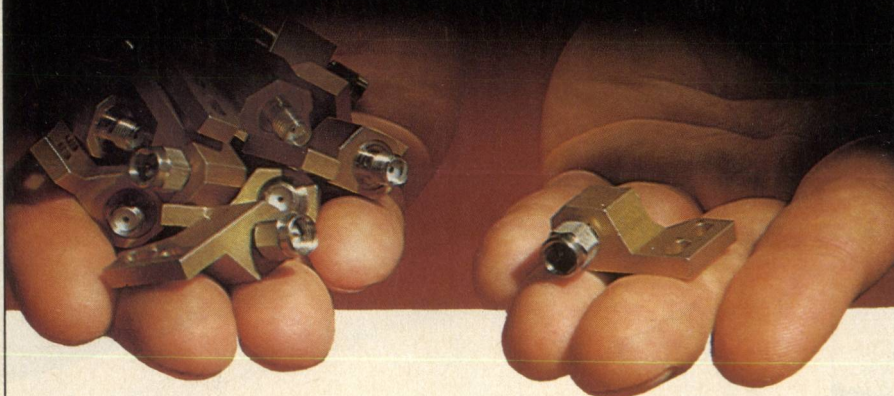


Fig. 11 Aerial view of the National Astronomy and Ionosphere Center.

[Continued on page 122]

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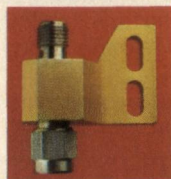
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[From page 120] **RADAR**

What Next Is Up To You

The Arecibo facility, The Natural Astronomy and Ionosphere Center, (Figure 11) which is managed by Cornell University is funded at \$5M per year by the National Science Foundation (NSF) and with \$300,000 per year from NASA. The facility is available for the use of scientists who submit proposals for the intended use of the instrument. Past discoveries at Arecibo include:

- In 1965 that the Planet Mercury rotation is not locked to its orbital period about the sun.
- The mapping of the visible surface of the Moon to 10 km resolutions using the UHF radar.
- That the largest moons of Jupiter (400 million miles from earth) were covered with fields of broken ice, a fact later verified by the Voyager spacecraft.
- That Venus rotates backwards once every 243.01 earth days, may have volcanoes the size of Oklahoma and plateaus as large as Hudson Bay.
- That the Crab Nebular Pulsar slows its pulse by 36 nanoseconds per day, a pulsar believed to consist of a neutron star whose density is 1 million tons per cubic inch.
- Support for Einstein's theory of gravitational waves, based on the observation of a binary pulsar system over an 8 year period.
- That 3 particular quasars at the edge of the observable universe are 15 billion light years distant and moving away at nearly the speed of light.

What might be discovered next? Perhaps that will be determined by a proposal that you write. I look forward to reading about your discovery.

References

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Passive Microwave Thermography Coupled With Microwave Heating To Enhance Early Detection Of Cancer

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Introduction

It is well known that a carcinoma, or malignant tumor, is normally hotter than the surrounding tissue and that, from "black body" theory, any perfectly absorbing body emits radiation at all frequencies in accordance with Planck's radiation law. The distribution of radiation is a function of both the temperature and the wavelength, or frequency. It should be noted that as the temperature of the body increases, the density of the radiation or emission at all frequencies also increases. From this viewpoint, infrared thermography, or radiometry, would appear effective; however, transmission losses increase with increasing frequency. Although the highest values of emission occur at the infrared region, an appreciable level occurs at the microwave region.

It is further known⁴ that the tumor tissue will die at temperatures above 42°C, and it has been reported⁵ that tumor temperatures of greater than 45°C can be held with adjacent normal tissue remaining at or near normal temperature.

The amount of microwave energy needed to provide localized heating is surprisingly low. If we neglect transmission losses and assume that tumor tissue has high water content, similar to that of muscle tissue, and further note that due to vascular insufficiency the tumor will retain and not readily dissipate the absorbed heat, then the amount of power required to elevate the temperature of a tumor 2 cm in diameter, 5° in 1

minute, is less than 1.5W. This can be shown as follows:

Quantity of heat,
 $Q = MC\Delta t$, calories (1)
where mass, M = volume $\times$ specific gravity; specific gravity of water is 1 g/cm³; specific heat, °C, is 1 calorie/g °C; and change in temperature, ΔT , = 5°C.

The energy Q required to elevate the temperature of the tumor 5°C is 20.95 cal. The power, P , therefore to effect this change in 1 minute will be

$$P = \frac{4.18Q}{t}, \text{ or } 1.46W \quad (2)$$

where 1 cal = 4.18 J and 1W = 1 J/S. This calculation assumes that

all the heat is transferred to the tumor and the tumor, in turn, does not dissipate this heat. Obviously, if the tumor is near the surface, the actual power required will approximate the calculated value, whereas with a deep tumor transmission losses, scattering, etc., will necessitate additional transmitter power.

This paper describes the design approach taken in the development of the microwave system, shown in Figure 1, to diagnose and treat cancer using non-invasive techniques. A sensitive passive microwave radiometer specifically designed to sense subsurface temperatures is coupled with a solid-state transmitter to provide localized heating of subsurface tissue, thereby taking advantage of the differential heating due to vascular insufficiency associated with the thermal characteristics of tumors to enhance early detection of cancer.

Design of Microwave System

The system is totally battery operated, allowing approximately 8 hours of continuous service prior to recharging. A battery-operated system totally eliminates possible problems associated with line transients, pickup, etc. The portable microwave system is contained in a standard 19" relay cabinet containing four drawers. The individual drawers are, in essence, totally enclosed housings to minimize interaction — particularly between the sensitive radiometer and the transmitter. The entire system is, in turn, mounted on a movable

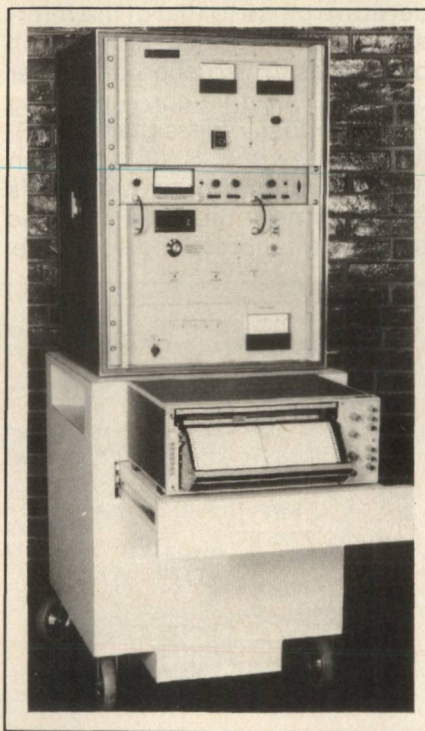


Fig. 1 Microwave system - drawer extended.

cabinet as shown in Figure 1.

The selection of both the radiometer and the transmitter frequencies was based upon several factors:

- Emissivity, which increases with increasing frequency;
- Spatial resolution, which increases with increasing frequency;
- Microwave transmission characteristics (microwave transmission losses generally increase with increasing frequency); and
- Microwave interference

Based upon the above factors, the frequency chosen for the microwave radiometer was 4.7 GHz which is far removed from the microwave heating frequency of 1.6 GHz.

The microwave radiometer is of the common load comparison, or Dicke, configuration. This configuration greatly reduces the effects of short-term gain fluctuations on the radiometer. The receiver input is switched at a constant rate between the antenna and a constant temperature reference load. The switched, or modulated, RF signal is, therefore, inserted at a point prior to RF amplification and as close to the antenna as possible. In turn, it is then amplified and coherently detected. The final output is not proportional to the temperature difference between the antenna and reference load. If a second switch (SW1 of Figure 2) is added, the reference load can now be compared with the base load rather than the antenna. If the base load is equal in temperature to the reference load, the DC output of the radiometer will be zero; i.e., a null will be achieved.

In the case where long integration times are involved, long-term gain variations in the receiver must be considered. The long-term, or slow, gain variations can degrade the minimum detectable temperature sensitivity (ΔT) in accordance with the following expression:

$$\Delta T \text{ (variation due to long-term gain change)} = \frac{\Delta G}{G} |T_1 - T_2|, ^\circ K$$

where ΔG is the receiver gain change, G is the nominal receiver gain, T_1 is the temperature of reference load in $^\circ K$, T_2 is the temperature of base load or antenna in $^\circ K$ (function of calibration switch position).

Obviously, if T_1 approaches T_2 , the effect of long-term receiver gain variations becomes negligible. It becomes advantageous, therefore, to maintain the temperature of both the base load and the reference load approximately equal to the temperature of the antenna.

The radiometer design employs a low-noise RF amplifier in conjunction with a simple single-ended square-law detector rather than the common super-heterodyne configuration involving a local oscillator and IF amplifier, thereby minimizing the potential drift and noise associated with this approach.

Figure 2 represents the simplified block diagram of the radiometer. The calibration switch, SW1, is a solenoid-operated mechanical single-pole double-throw switch used to disconnect the antenna and, in turn, to connect the comparison base load. The base load is maintained at a temperature close to that of the reference load. The mechanical switch possesses an isolation of greater than 60 dB with a corresponding insertion loss of less than 0.1 dB. A similar switch, SW2, is used in the calibrate path. Its function is to disconnect the base load and to insert a calibrated noise source.

As shown in Figure 2, there are three ferrite isolators used in the receive path. The first isolator, located between the calibration switch and the Dicke switch, is used to terminate the output of the reference load when the Dicke switch is in the low loss state. In this state, the reference load is circulated in the direction of the antenna which, in this case, is the ferrite isolator. The second isolator, which is located between the Dicke switch and the first-stage RF amplifier, maintains a constant load match to the amplifier. Any reflections from the RF amplifier would, therefore, be terminated in the isolator. A third isolator is located between the bandpass filter and the output of the first RF amplifier. The purpose of this particular isolator is to present a matched input to the bandpass filter.

A switchable ferrite circulator, SW3, has been developed to perform the load comparison or Dicke switch function. The ferrite switch is preferred to the semiconductor approach primarily in view of the lower insertion loss, typically less than 0.25 dB, and elimination of noise generated by the semiconductor junction over and above the measured insertion loss. Briefly, the device is a switchable ferrite junction circulator utilizing the remnant or latching characteristics of the ferrite material. The latching ferrite switch has been constructed in a waveguide having a single ferrite element contained within the microwave circuit. The insertion loss was

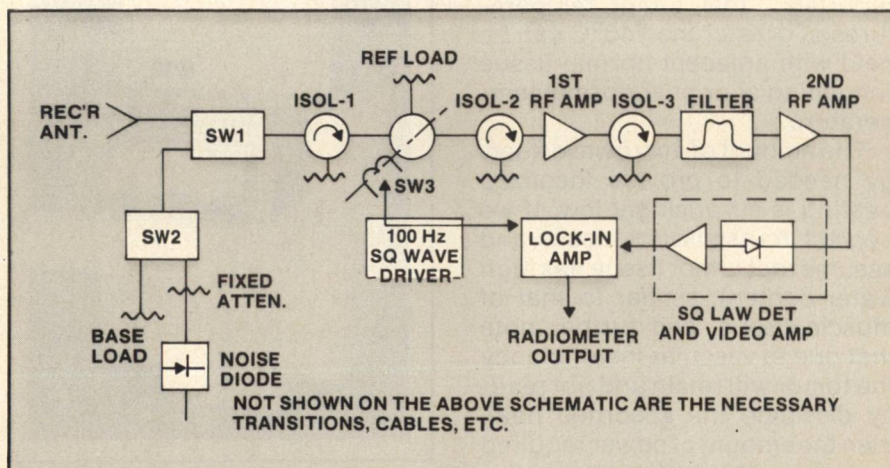
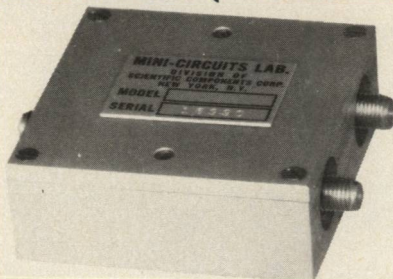


Fig. 2 Microwave radiometer schematic.

[Continued on page 128]

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ZAM-42 SPECIFICATIONS

FREQUENCY RANGE, (GHz)			
LO, RF	1.5-4.2		
IF	DC-0.5		
CONVERSION LOSS, dB			
Total range	TYP.	MAX.	
	7.0	8.5	
ISOLATION, dB			
	TYP.	MIN.	
1.5-2.0 GHz LO-RF	25	20	
LO-IF	18	10	
2.0-3.7 GHz LO-RF	25	17	
LO-IF	18	10	
3.7-4.2 GHz LO-RF	25	20	
LO-IF	18	10	

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[From page 126] THERMOGRAPHY

measured and found to be less than 0.25 dB, having isolation in excess of 20 dB.

The RF amplifier is a four-stage FET device constructed in microstrip with integrated biasing circuitry. The noise figure of the first RF amplifier is 2.2 dB with a gain of 35 dB. The second RF amplifier has a noise figure of 2.6 dB with an associated gain of 33 dB. In both instances, the noise figure includes the input ferrite isolator as shown in Figure 2. With the input and output VSWR at less than 1.5:1, the gain compression at signal levels of between -55 to -10 dbm was less than 0.1 dB.

The bandwidth of the microwave radiometer is basically determined by the bandpass characteristics of the filter. The filter is located after the first RF amplifier to minimize the impact of the insertion loss on the overall system performance. The filter characteristics were chosen to minimize possible interference due to nearby microwave communications on radar bands. The printed, eight-section, bandpass filter is constructed in stripline. The bandwidth is approximately 500 MHz centered at 4.7 GHz. The insertion loss in the passband is less than 3 dB.

The lock-in amplifier enables the accurate measurement of signals contaminated by broadband noise, power-line pickup, frequency drift, or other sources of interference. It does this by means of an extremely narrowband detector which has the center of its passband locked to the frequency of the signal to be measured. Because of the frequency lock and the narrow bandwidth, large

improvements in signal-to-noise ratio can be achieved. This allows the signal of interest to be accurately measured, even in situations where it is completely masked by noise. In addition, the lock-in amplifier provides the synchronous function associated with the Dicke switch; i.e., the unit supplies the 100 Hz reference clock frequency to drive the ferrite switch driver.

The minimum detectable temperature sensitivity ΔT was well within the desired 0.1°C. The minimum detectable temperature sensitivity ΔT is expressed as follows:

$$\Delta T = \frac{k [(FL-1) T_1 + T_2]}{\sqrt{B\tau}}, \text{ } ^\circ\text{K. (4)}$$

In the case of the Dicke switch employing square-wave modulation, the value of k is 2.0; F is the noise figure (first amplifier stage), which in our case is 2.2 dB (1.66 ratio); and L is the input losses, expressed as a power ratio. The total loss is 2.0 dB (1.58 ratio).

The effective noise figure FL is, therefore, 2.2+2, or 4.2, which represents a power ratio of 2.63. T_1 is the ambient radiometer temperature (microwave portion); namely, 290°K; T_2 is the source temperature (i.e., temperature seen by antenna), namely 310°K; B is the receiver bandwidth (i.e., the 3 dB bandwidth of the bandpass filter following the first RF amplifier), namely, 500 MHz; and τ is the radiometer output time constant, in seconds.

Utilizing a 3-s time constant, we obtain a minimum detectable temperature sensitivity of

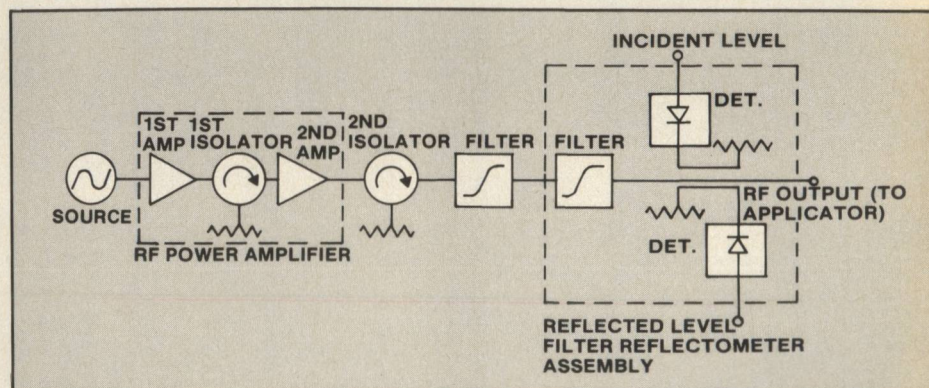


Fig. 3 L-Band transmitter schematic.

$$\Delta T = \frac{[2 (2.63-1) 290+310]}{\sqrt{500 \times 10^6 \times 3}}, (5)$$

or 0.04°K rms.

Increasing the time constant τ to 10 s results in a ΔT of 0.02°K. Similarly, reducing the time constant to 1 s will result in a ΔT of 0.07°K. The calculated temperature sensitivities are well within our design goal of 0.1°C, as were the actual results.

Microwave Transmitter

The microwave transmitter, shown schematically in Figure 3, consists of a 1.6 GHz 30W solid state source followed by a low pass filter and microwave reflectometer. Sufficient filtering has been provided through the use of two low pass filters in series providing 120 dB of attenuation at the third harmonic.

The third harmonic of the 1.6 GHz source is 4.8 GHz which is well within the radiometer pass-band. The reflectometer, in turn, allows determination of both the reflected and incident power levels. The output power level is electronically adjustable from 0 to 25W measured at the output of the L-band antenna and, therefore, includes all microwave circuit and coaxial cable losses.

Dual Mode Applicator

It has been determined by Guy⁶ that the optimum aperture size to achieve effective coupling of microwave energy associated with layered biological tissue is the simple TE₁₀ mode aperture in direct contact with the radiated, or emitting, surface. A normal waveguide transition at L-band corresponding to WR-510 would be 5.10" (12.95 cm) x 2.55" (6.48 cm). To reduce the physical size of this aperture, single-ridged waveguide was employed since the use of ridged waveguide lowers the frequency at which cutoff will occur. To further reduce the overall size of the aperture, dielectric loading was used.

The dimensions of the ridged portion of the L-band ridged waveguide were selected to allow propagation of the higher frequency associated with the C-band radiometer. The plated surface of the

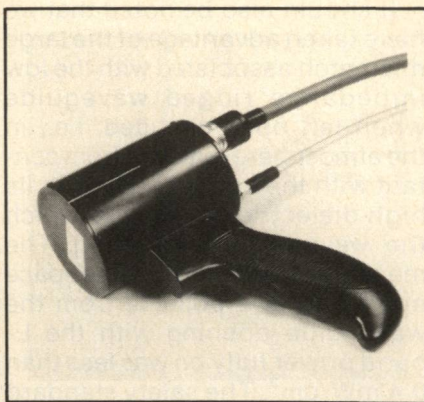


Fig. 4a Dual mode applicator

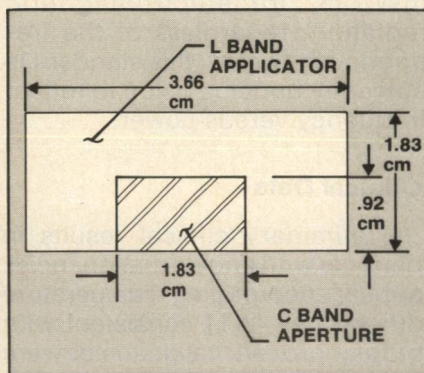


Fig. 4b Dual mode applicator face

dielectric-loaded C-band waveguide, therefore, formed the single ridge of the L-band waveguide as shown in Figure 4. The dielectric material used is aluminum oxide having a dielectric constant, ϵ_r , of 9.8. The C-band aperture dimensions are 0.92 cm x 1.83 cm. The overall dimension of the L-band waveguide is 3.66 cm x 1.83 cm.

It can be seen in the aforementioned figure that by having the radiometer input contained within the single-ridged waveguide L-band transition, we have placed the point of maximum field of the source of heat in close proximity with that point of thermal detection. The cutoff characteristics of the C-band waveguide have been utilized in addition to other filtering provided, the waveguide forming a high pass filter to isolate the high power L-band source from the sensitive radiometer. A heater and proportional thermostat are provided in the dual mode transition, or antenna, to maintain a constant temperature at or very near to that of the temperature of the human body.

[Continued on page 130]

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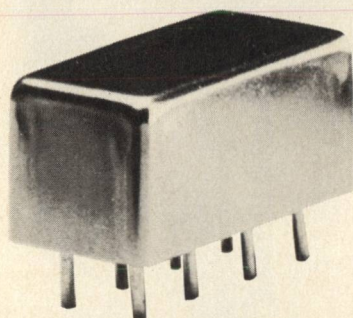


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FREQUENCY (MHz) 0.2-250
COUPLING, db 19.5

INSERTION LOSS, dB	TYP.	MAX.
one octave band edge	0.35	0.5
total range	0.35	0.6

DIRECTIVITY, dB	TYP.	MIN.
low range	36	30
mid range	32	25
upper range	25	20

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[From page 129] THERMOGRAPHY

It should also be noted that we have taken advantage of the large mismatch associated with the low impedance ridged waveguide when left open-circuited; i.e., in the atmosphere removed from contact with the human body with its high dielectric constant to which the waveguide is matched. The measured power level in free space measured at a point 1" from the waveguide opening with the L-band power fully on was less than 0.4 mW/cm^2 . The safety standard established by ANSI has been 10 mW/cm^2 for electromagnetic radiation, regardless of the frequency; however, this standard is currently under revision to reflect frequency versus power.

Clinical Data

Preliminary clinical results in patients with known malignancies appear encouraging. Temperature differences (ΔT) consistent with biopsy-proven malignancies were obtained in 4 of 6 primary carcinomas of the breast (Table I), in 5 of 5 patients with lymphoma (Table II), and in 10 of 12 women with recurrent breast cancer (Table III). Negative results were obtained mainly in patients with more deeply-seated tumors. It should be noted (Table I) that one breast cancer patient (79-001) had

bilateral tumors; therefore, there were a total of 4 tumors in the 6 patients.

It should be pointed out that the above referenced results were obtained using passive detection alone and did not involve the use of microwave heating. It has been reported⁸ that tumor temperatures can be elevated with respect to surrounding normal tissue. The use of localized heating, therefore, should provide a technique to highlight the tumor... thereby enhancing detection. Attempts to date to locate more deeply-seated (lung, esophagus, femur, humerus) malignancies by microwave thermography alone have been unsuccessful. From a theoretical standpoint, if the temperature differential between the tumor and adjacent normal tissue could be increased, one might reasonably expect to be able to detect smaller and/or more deeply-seated tumors.

It is noteworthy (Table I) that although negative results were obtained in 2 breast cancer patients (79-009) and (80-022) for whom xeromammograms were positive, the microwave thermograms were positive in 2 patients (79-011) and (80-016) whose xeromammograms were negative or, at best, inconclusive. It should likewise be mentioned that in the

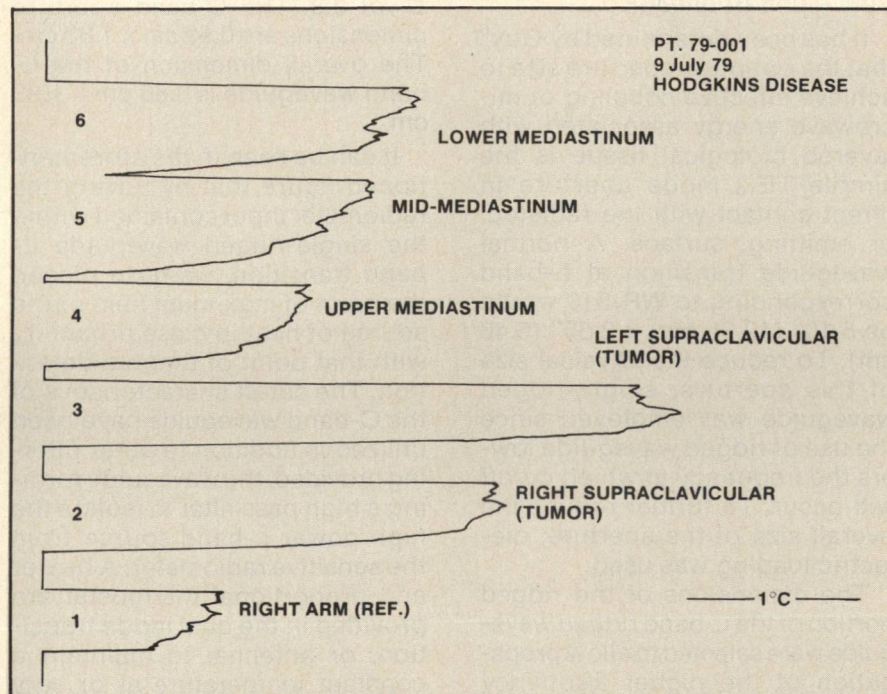


Fig. 5 Pre X-ray therapy thermogram

[Continued on page 132]
MICROWAVE JOURNAL

patient with bilateral breast cancer (79-011), the larger ΔT value of 2.5°C was observed in the smaller lesion (<10mm diameter) in the left breast which was not seen in the xeromammograms. The smaller ΔT value of 1.2°C was obtained in the right breast lesion of approximately 2.5 cm diameter, which was seen on xeromammographic (X-ray) studies.

Patient 79-001 presented with nodular sclerosing Hodgkin's disease in her right and left supraclavicular areas (Figure 5, peaks 2 and 3). This examination was done prior to X-ray therapy. Several weeks later following a course of radiation therapy (4000 Rads), there was complete resolution of the tumor with a concomitant reduction in the ΔT (Figure 6). Like-

wise there was an increase in ΔT in one patient (79-012), Table III) whose tumor did not respond to radiation therapy. Thus, microwave thermography may have some potential in assessing the effectiveness of various therapeutic modalities. Placement of the antenna on the right forearm was used to establish a reference temperature.

TABLE I
CLINICAL MICROWAVE THERMOGRAMS — PRIMARY CARCINOMA OF THE BREAST

PATIENT	AGE	SEX	DIAGNOSIS	SITE	ΔT
79-009	58	F	Adenocarcinoma/Breast	Right Upper Outer Quadrant	0
79-011	47	F	Adenocarcinoma/Breast	Right Upper Outer Quadrant	1.2
				Left Upper Inner Quadrant	2.5
80-016	64	F	Adenocarcinoma/Breast	Left Upper Inner Quadrant	1.1
80-022	73	F	Adenocarcinoma/Breast	Right Upper Outer Quadrant	0
80-024	33	F	Inflammatory Carcinoma/Breast	Right Breast	1.3

TABLE II
CLINICAL MICROWAVE THERMOGRAMS — LYMPHOMA

PATIENT	AGE	SEX	DIAGNOSIS	SITE	ΔT
79-003	17	M	Hodgkin's Disease	Mediastinum	1.2
79-001	26	F	Hodgkin's Disease	Right Supraclavicular	1.6
80-025	31	M	Hodgkin's Disease	Right Supraclavicular	0.9
80-030	58	M	Giant Follicular	Right Shoulder	1.6
			Lymphoblastoma		
80-	70	F	Histiocytic Lymphoma	Right Leg	0.4*

*Denotes ΔT in an unresponsive tumor following a course of radiation therapy.

TABLE III
CLINICAL MICROWAVE THERMOGRAMS — RECURRENT BREAST CANCER

PATIENT	AGE	SEX	DIAGNOSIS	SITE	ΔT
80-017	34	F	Recurrent Breast Cancer	Left Supraclavicular	0
79-006	51	F	Recurrent Breast Cancer	Left Anterior Chest	3.3
79-013	57	F	Recurrent Breast Cancer	Left Anterior Chest	2.2
79-012	76	F	Recurrent Breast Cancer	Right Anterior Chest	1.1
79-002	54	F	Recurrent Breast Cancer	Right Anterior Chest	1.6
80-018	75	F	Recurrent Breast Cancer	Left Anterior Chest	2.5
80-029	83	F	Recurrent Breast Cancer	Left Anterior Chest	0.9
80-031	49	F	Recurrent Breast Cancer	Left Anterior Chest	0.8
80-034	50	F	Recurrent Breast Cancer	Left Anterior Chest	2.5
80-037	57	F	Recurrent Breast Cancer	Anterior Chest-bilateral	2.1

KRYTAR...

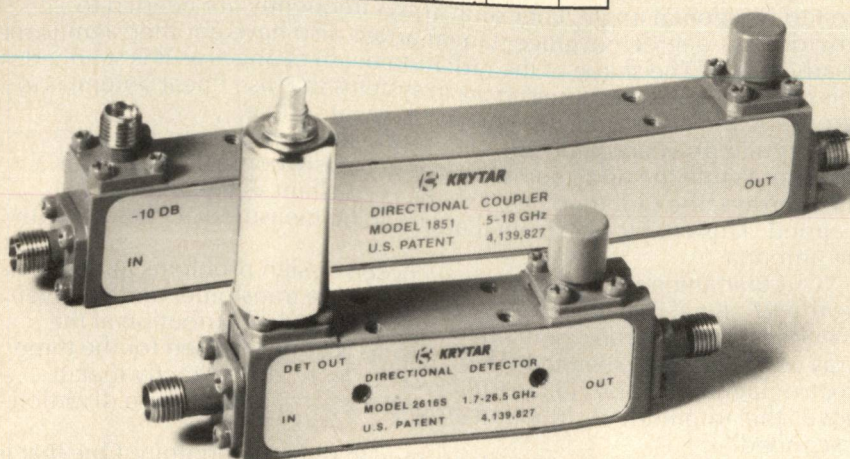
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1211S	1-12.4	$\pm 2(1-8)$ $\pm 3(1-12.4)$	18(1-8) 15(8-12.4)	1.35	1.1	40	\$675
1818S	2-18	$\pm 5(2-12.4)$ $\pm 7(2-18)$	17(2-12.4) 15(12.4-18)	1.35	.75	10	\$750
1822S	2-18	$\pm 5(2-12.4)$ $\pm 7(2-18)$	15(2-12.4) 13(12.4-18)	1.35	1.0	40	\$750
1820S	1-18	$\pm 5(1-12.4)$ $\pm 7(1-18)$	17(1-12.4) 15(12.4-18)	1.35	.9	10	\$825
1821S	1-18	$\pm 5(1-12.4)$ $\pm 7(1-18)$	15(1-12.4) 13(12.4-18)	1.40	1.2	40	\$825
1850S	.5-18	± 1.2	14(.5-18) 12(12.4-18)	1.40	1.1	10	\$925
1851S	.5-18	± 1.2	14(.5-12.4) 12(12.4-18)	1.40	1.5	40	\$925
2616S	1.7-26.5	$\pm 8(1.7-18)$ $\pm 1.2(1.7-26.5)$	15(1.7-18) 13(18-26.5)	1.45	1.2	10	\$1,125
3617S	3.6-11.7	$\pm 15(3.6-6.5)$ $\pm 30(6.5-11.7)$	18(3.6-6.5) 15(6.5-11.7)	1.30	1.1	40	\$825

DIRECTIONAL COUPLERS

Model	Frequency (GHz)	Nominal Coupling (dB)	Frequency Sensitivity (dB) (GHz)	Directivity (dB) (GHz)	Max VSWR	Insertion Loss (dB Max)	Price
1211	1-12.4	10 $\pm$.5	± 2	18(1-8) 15(8-12.4)	1.35	1.1	\$475
1818	2-18	16 $\pm$.5	$\pm 25(2-12.4)$ $\pm 35(2-18)$	17(2-12.4) 15(12.4-18)	1.30	.75	\$475
1822	2-18	10 $\pm$.5	$\pm 25(2-12.4)$ $\pm 4(2-18)$	15(2-12.4) 13(12.4-18)	1.35	1.0	\$475
1820	1-18	16 $\pm$.5	$\pm 3(1-12.4)$ $\pm 4(1-18)$	17(1-12.4) 15(12.4-18)	1.35	.9	\$475
1821	1-18	10 $\pm$.5	± 4	15(1-12.4) 13(12.4-18)	1.40	1.2	\$475
1850	.5-18	16 $\pm$ 1	± 1	14(.5-12.4) 12(12.4-18)	1.35	1.1	\$575
1851	.5-18	10 $\pm$ 1	± 1	14(.5-12.4) 12(12.4-18)	1.35	1.5	\$575
2616	1.7-26.5	16 $\pm$ 1	$\pm 4(1.7-18)$ $\pm 8(18-26.5)$	15(1.7-18) 13(18-26.5)	1.45	1.2	\$725



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[From page 132] THERMOGRAPHY

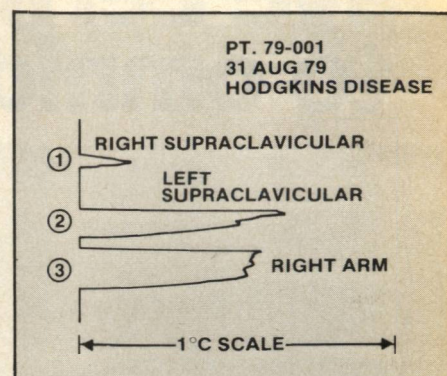


Fig. 6 Post X-ray therapy thermogram

Insofar as mass screening of breast cancer is concerned, microwave thermography has several distinct advantages over X-ray xeromammography. First, there is no known risk to the patient since the unit is merely monitoring the patient's natural microwave emissions; no harm is done to the patient. Second, there is a possibility (which must be borne out through experimentation) that microwave thermography may be able to detect smaller tumors that can be seen by conventional techniques. This statement has great therapeutic significance since the probability of achieving a cure is inversely related to the size of the tumor at diagnosis.

Acknowledgement

The authors wish to acknowledge the assistance of T. Felderman, M.D. and Messrs. R. Bielawa and J. Regan in this investigation.

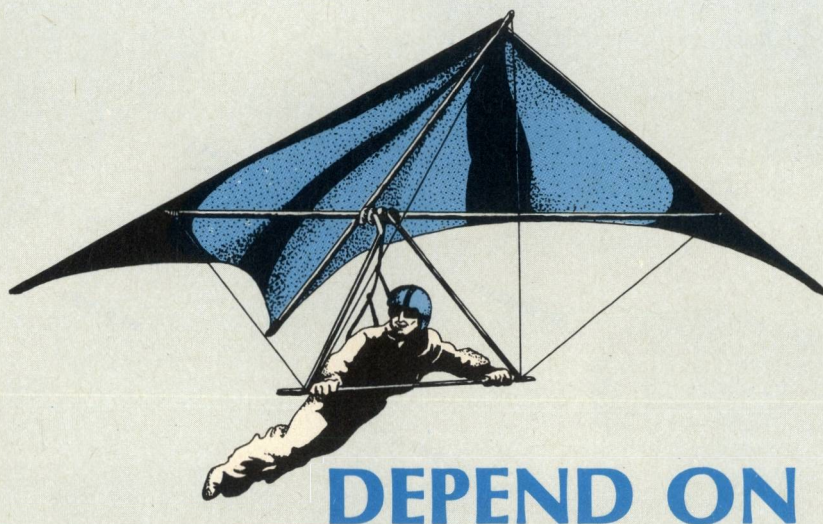
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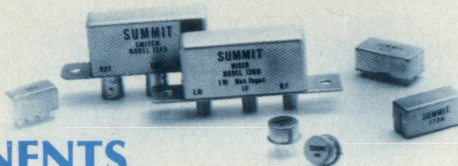
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[From page 134] **THERMOGRAPHY**

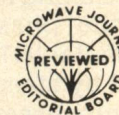
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Programmable SAW Transversal Filters

Richard T. Webster

Electromagnetic Sciences Division
Rome Air Development Center
Hanscom AFB

Introduction

Surface acoustic wave (SAW) technology allows fabrication of compact tapped delay lines which have large time-bandwidth products, wide dynamic range, and low spurious levels. Extremely versatile electronically programmable transversal filters (PTF) can be obtained by integrating these tapped delay lines with control electronics. These filters find numerous applications in radar and spread spectrum communication systems for adaptive frequency filtering and for asynchronous correlation of coded waveforms. This report examines several approaches to obtaining PTF's using SAW tapped delay lines. Both hybrid and monolithic techniques are discussed and examples are given to illustrate the capabilities, the advantages and the limitations of each approach.

Figure 1 shows a basic transversal filter configuration where a signal propagates through a series of equal delays of duration T . Following each delay interval, the signal is sampled, then the samples are weighted in both amplitude and phase. Finally, all the tap signals are summed to form the filter output. After undergoing the delay and weight process, some frequency components of the signal add in phase while others do not. Hence, the transversal filter is frequency selective. The sequence of tap weights forms the filter impulse response. Consequently, the Fourier transform of the tap weights is the frequency response of the filter. Electronically programmable tap weights enable tailoring of the frequency response within the bandwidth limitation established by the re-

ciprocal of the intertap delay time. The addition of the appropriate tap weight control circuitry would create an adaptive filter for noise cancellation, frequency tuning, multipath suppression, or adaptive equalization of channels when the precise channel characteristics are unknown.

Another important use of the tapped delay line transversal filter is for asynchronous correlation or matched filtering of spread spectrum waveforms. In a matched filter, the signal-to-noise ratio is maximized when the impulse response of the filter equals the time reverse of the signal waveform. This function is employed for transmitting signals over noisy channels and for covert transmission where only those receivers having matched filters can detect the message. For the latter application, it is desirable to change the code frequently to prevent unauthorized receivers from deciphering the message. A programmable filter is therefore required so the impulse response can be

changed to match the changing code.

SAW technology is naturally suited to perform transversal filtering for the following reasons. First, useful signal delays can be realized in relatively small packages because the SAW velocity is roughly a factor of 10^{-5} slower than the velocity of electromagnetic radiation. Delays of several microseconds occur in dimensions of the order of centimeters. Second, since the wave propagates along the surface, suitable transducers can detect it at any point along the propagation path. On a piezoelectric substrate, the interdigital transducer (IDT) is a very effective detector. In fact, the IDT itself is a form of transversal filter. The transducer electrodes tap the signal as it propagates through the IDT. The bus bars or bonding pads perform the summing. Amplitude weighting can be accomplished in several ways, the most common of which are electrode length apodization and electrode withdrawal weighting.

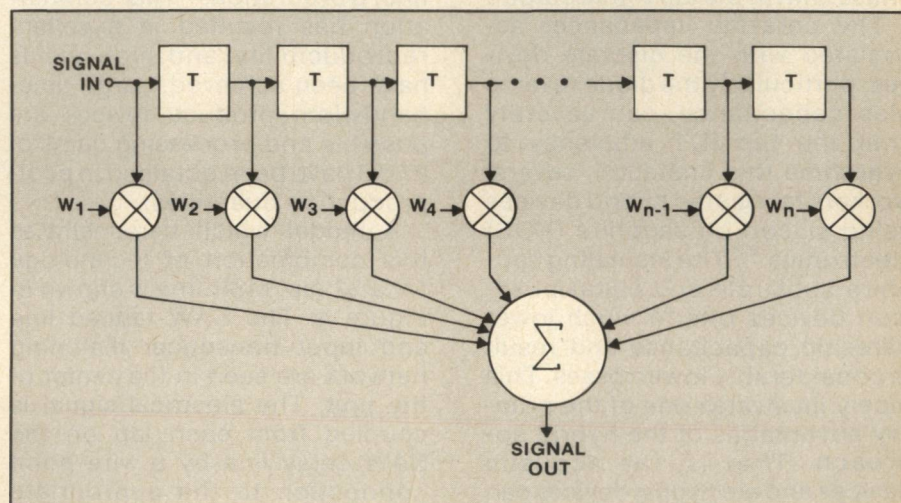


Fig. 1 Schematic diagram of a transversal filter.

The relative positions of the electrodes along the delay path establish the phase weighting. Thin film metal IDT's are readily fabricated by standard photolithographic processes and detailed transducer design techniques have been developed resulting in accurate fixed filter responses in compact, rugged packages. These attractive features contribute to the widespread acceptance of fixed response SAW bandpass and matched filters. We now examine how many of these desirable features can be retained in devices which have the added flexibility of a programmable filter response.

The Hybrid Approach

The straight-forward approach to achieving programmability is to choose a standard piezoelectric SAW substrate, fabricate a series of IDT's in a tapped delay line configuration and wirebond each tap IDT to nearby weighting, summing, and control electronics as shown in Figure 2.

One early example of the hybrid approach demonstrated a biphasic programmable matched filter by using discrete PIN diodes in an SPDT switch configuration like that shown in Figure 3.¹ A similar switch would be connected to each tap. A positive programming voltage, V_p , forward biases diodes D1 and D4 while reverse biasing D2 and D3. The left tap electrode is thus connected to the ground bus and the right electrode to the output bus. Reversing the polarity of the programming voltage, V_p , reverses the electrode connections producing a 180° phase shift in the tap signal output.

The parasitic impedances associated with the discrete devices, particularly the diode reverse bias capacitance, can severely limit the tap IDT efficiency to overcome this limitation, several workers fabricated hybrid devices using silicon on sapphire (SOS) electronics.^{2,3} The insulating sapphire substrate and epitaxial silicon devices offer a much lower parasitic capacitance and result in considerably lower losses. This nicely illustrates one of the primary advantages of the hybrid approach. That is, the acoustic devices and electronic devices can be independently optimized.

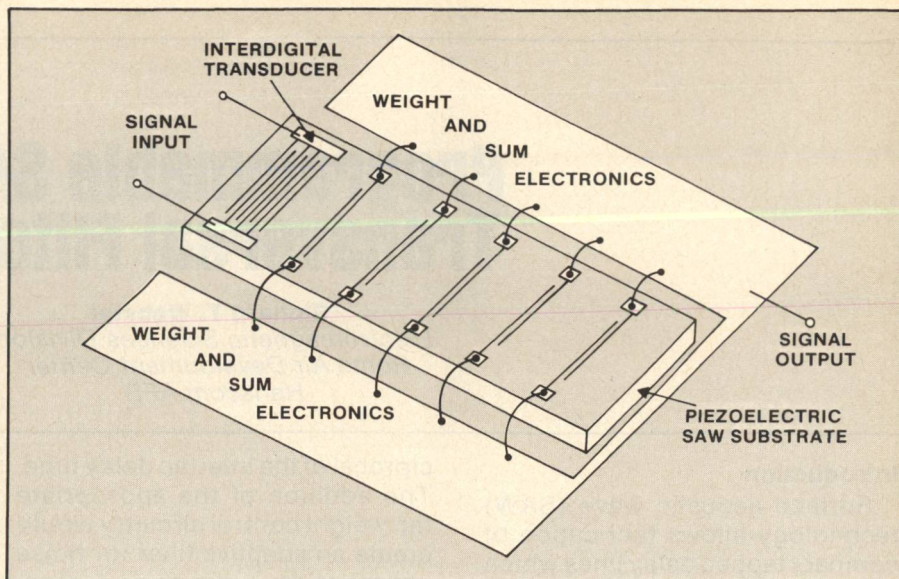


Fig. 2 The hybrid approach to SAW PTF's uses a tapped delay line on a standard SAW substrate with wire bonds connecting the taps to electronic circuitry.

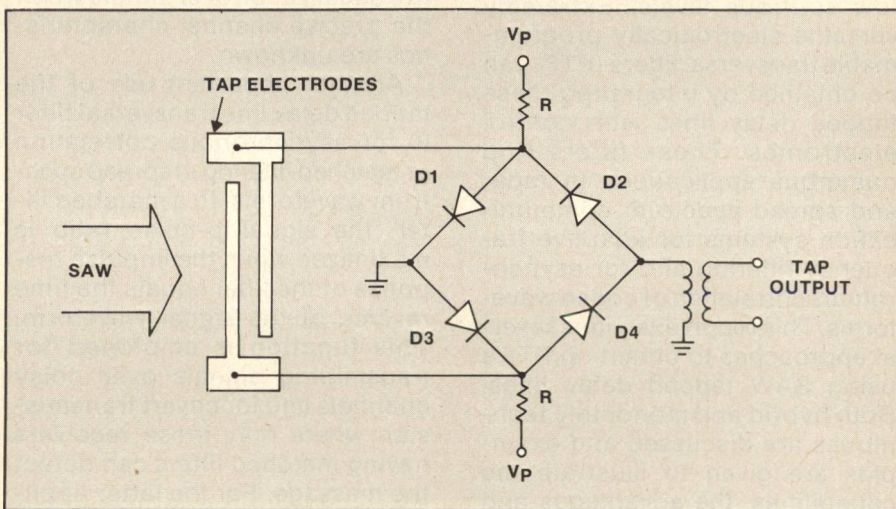


Fig. 3 An SPDT diode switch can be implemented for biphasic programmability.

Hybrid programmable matched filters which combine quartz delay lines and thick film technology have been developed by Hazeltine Corporation.⁴ This combination has resulted in excellent reproducibility and high yields have been achieved. Large time-bandwidth product devices are possible and processing gains of 27 dB have been obtained in pilot-production quantities.

A model which demonstrates this combination of technology for a 32 tap, MSK line is shown in Figure 4. The SAW tapped line and input transducer matching network are seen in the center of the unit. The electrical signal is coupled from each tap on the SAW delay line by a wire bond connection to the appropriate thick film hybrid substrate. Each

substrate contains the RF switching and logic control circuits necessary to process the PSK, QPSK or MSK spread spectrum

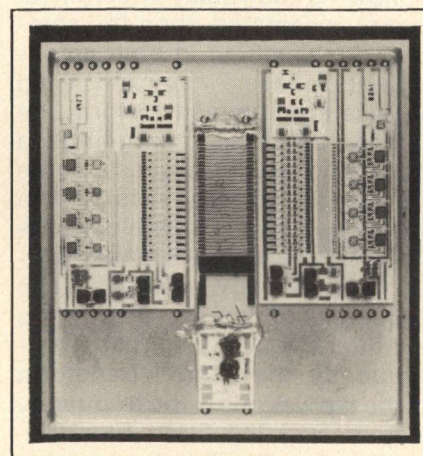


Fig. 4 A bit programmable tapped delay line illustrating combined SAW and thick film hybrid technology.

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(for 0 dBm test level)	
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[From page 140] FILTERS

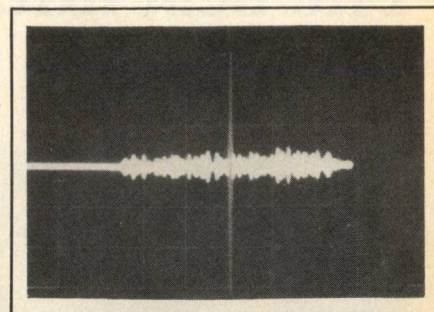


Fig. 5 Correlated output for 64-bit PSK input signal.

signal. The units operate satisfactorily from -55° to $+128^{\circ}\text{C}$ and meet military environmental specifications. After assembly, the case and cover are seam welded together in a dry nitrogen atmosphere since hermeticity is essential to ensure a contamination free environment for both the SAW line and semiconductor chips.

The correlated output pulse for a 64 bit programmable MSK device is shown in Figure 5. This device operates at a 75 MHz center frequency with a 5 MHz chip rate. The sidelobe level is within 2 dB, and the processing gain; within 0.5 dB of theoretical.

Additional size and cost reduction has been demonstrated through the use of custom LSI circuits to replace discrete components used in earlier models. In particular, a 25 bit bipolar RF switch and a 25 bit CMOS logic chip were developed to process the outputs of 25 contiguous taps. These custom LSI chips permit 128 bit control circuits to be realized in approximately the same volume as the earlier 32-bit control circuits. Currently under development is a 32-bit combined RF switching, summing and logic control chip which will effect an even further decrease in size and a lower cost in production.

Hybrid programmable filters having both amplitude and phase weighting have also been demonstrated. One technique uses non-linear mixing of two SAWs propagating on the same path.^{5,6} Grudkowski and Reeder⁶ used a configuration similar to the simplified one shown in Figure 6. A wave generated by the input signal at frequency F_1 and a CW reference wave at frequency F_2 are both detected by the taps. Each

(Hazelton Corp. photo)

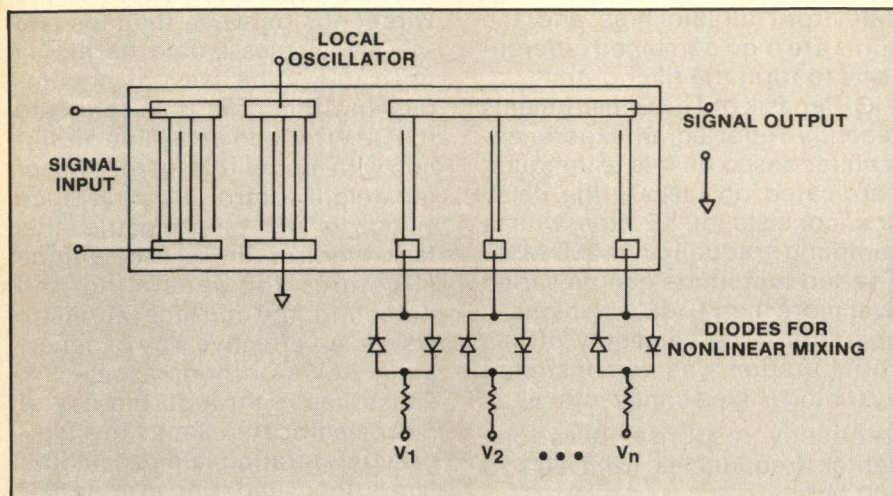


Fig. 6 Nonlinear mixing of tap signals in diode pairs allows both amplitude and biphase ($\pm$) tap weighting.

tap is bonded to an antiparallel pair of diodes fabricated as an array using SOS technology. The programming voltage applied to each tap forward biases one or the other of the diodes. The two detected signals mix in the forward biased diode producing product signals at the sum and difference frequencies, $F_2 + F_1$ and $F_2 - F_1$. The strength of the product signals is dependent on the level of the DC bias current. A phase shift of 180° occurs in the tap output signal when the direction of the DC current flow to the diode pair is reversed. An external bandpass filter at the output port selects the filtered signal at either the sum or difference frequency. A 128 tap filter was fabricated using a YZ-LiNbO₃ delay line and an SOS diode array. The nominal signal frequency was 70 MHz with the local oscillator at 100 MHz

and the output taken at the difference frequency. The tap sampling frequency was 9.8 MHz. A wide range of frequency responses was demonstrated including bandpass filters with bandwidths variable between 0.1 and 1 MHz. By varying the frequency of the local oscillator, the center frequency of the filter could be varied independently of the tap weights. Asynchronous correlation of a 127 chip M-sequence PSK code at a 10 MHz chip rate was also demonstrated.

Combining the output of two of these devices in a quadrature hybrid allowed the synthesis of complex tap weights. Impulse responses with arbitrary amplitude and phase characteristics within the time-bandwidth limitations of the device were synthesized by programming the tap weights of the real and imaginary channels.

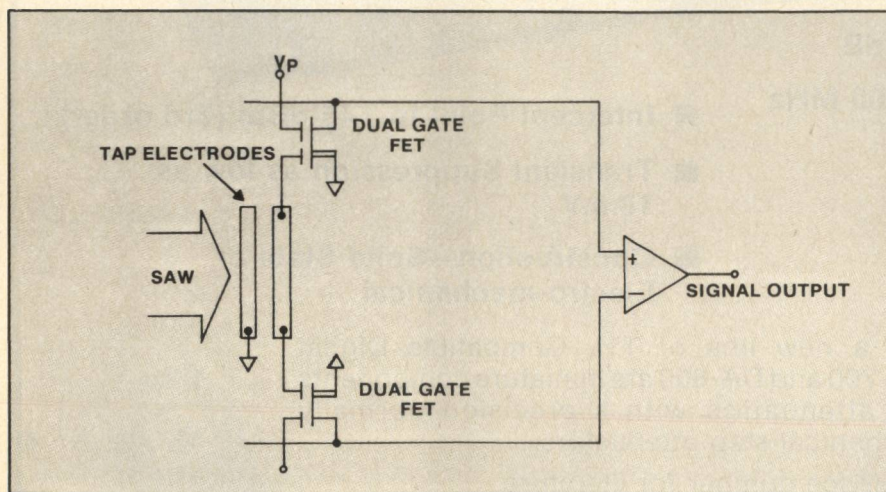


Fig. 7 A dual gate MOSFET configuration can achieve biphase tap weighting with linear amplitude control over more than 40 dB.

[Continued on page 144]

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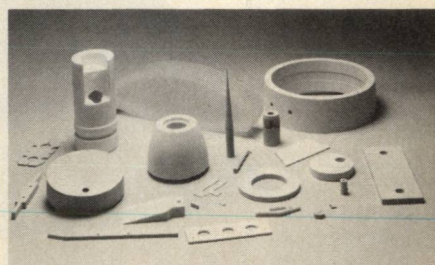
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This was graphically illustrated by the asynchronous correlation of linear chirp signals of various chirp rates.

Another tap weighting technique which is capable of both biphase and amplitude programmability was reported by Kwan, Chuang, Muller and White⁷. This method uses a pair of dual gate MOSFETS connected to each active tap electrode. Figure 7 shows the configuration for one tap. Bias circuitry is not shown. The voltage developed across the electrode as the SAW passes through is applied to the first gate of both MOSFETS. The programming voltage applied to the second gates control the strength of the signals which reach the two bus

bars. The bus bars sum the signals from all the taps and the sums are then combined differentially to form the filter output.

C. Panasik of Texas Instruments recently reported an experimental filter based on this technique.⁸ Fabricated on LiNbO₃, the delay line consists of 16 taps with a sampling frequency of 200 MHz. The tap amplitude can be varied over more than 40 dB with biphase operation. The versatility of this configuration was demonstrated by tuning a fixed bandwidth $\frac{\sin x}{x}$ frequency response to various center frequencies from 200 to 300 MHz.

The Monolithic Approach

On Nonpiezoelectric Substrates

For improved reliability and

ruggedness, elimination of the wirebonds between the taps and the electronics would be highly desirable. One way of accomplishing this goal is to fabricate the SAW tapped delay line on silicon with monolithic integration of tap weight control circuitry. Since silicon is not piezoelectric, this task requires the use of unique structures for generating and detecting the surface acoustic waves. An effective way of generating SAWs on nonpiezoelectric substrates is through the use of piezoelectric thin films.⁹ In a typical configuration, a piezoelectric film is deposited over an interdigital transducer. The acoustic wave is generated in the piezoelectric film and propagates as a SAW on

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the substrate. Considerable progress has been made in the development of zinc oxide ZnO thin film transducers. ZnO enjoys a relatively high piezoelectric coupling and continuing improvement of the sputter deposition technology is yielding reproducible, well-oriented films with piezoelectric properties approaching that of the bulk material. The process is compatible with standard silicon IC processing.

Propagation losses and dispersion effects make it difficult to fabricate an entire tapped delay line using ZnO on silicon. These problems have encouraged the use of a different phenomena for the programmable taps. This is the piezoresistance effect wherein stress waves propagating in silicon alter the effective mobility of charge carriers. If the waves propagate through a suitably oriented and properly biased MOSFET, the signal carried by the SAW appears as a modulation of the drain current.¹⁰

A fine example of this technology is by the work performed at

Motorola under the direction of F.S. Hickernell.¹¹ A recent effort produced a programmable correlator capable of processing 31 bit biphasic coded signals at a center frequency of 100 MHz, with a code rate of 10 MHz, and a duration of 3.1 microseconds. The resulting LSI chip includes a thin film transducer, biphasic MOSFET taps and integrated switching and logic control circuitry.

The device configuration is similar to that shown in Figure 8. The SAWs are generated by a thin film ZnO overlay transducer fabricated on glassivated silicon. They propagate on the (001) silicon surface in the [100] direction. An array of piezoresistive MOSFET taps lies in the propagation path. The sampling frequency is 10 MHz. Biphasic operation is achieved by using a split detector structure consisting of two MOSFETs sharing a common drain like the one shown schematically in Figure 8. The gate electrodes of this split detector tap are separated by a center-to-center distance equal to one-half the SAW wavelength.

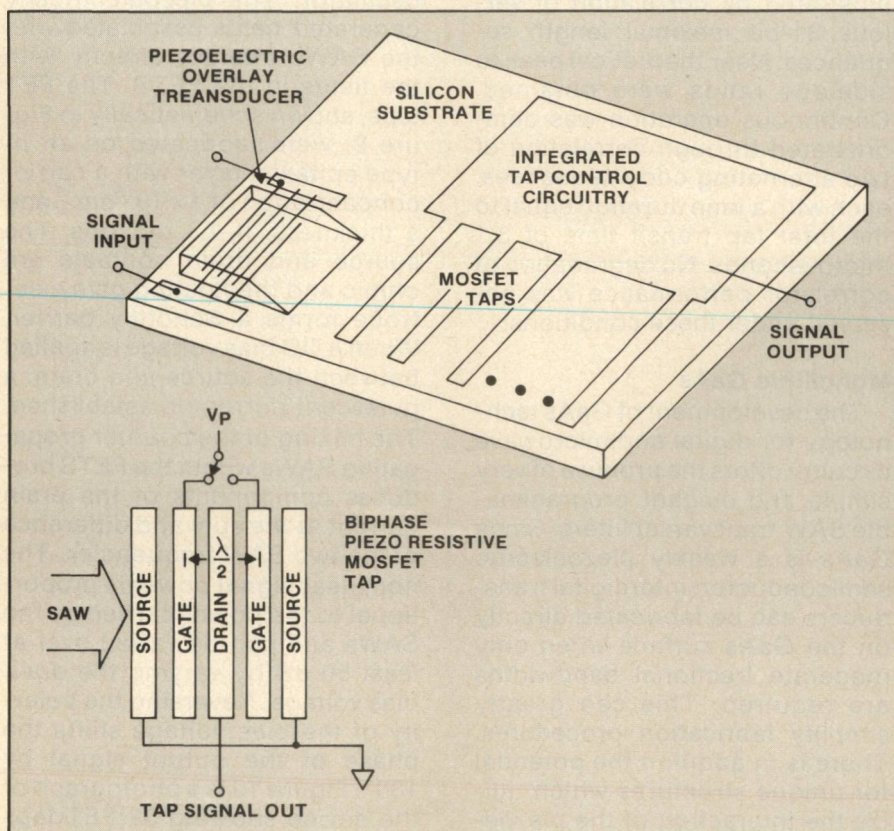
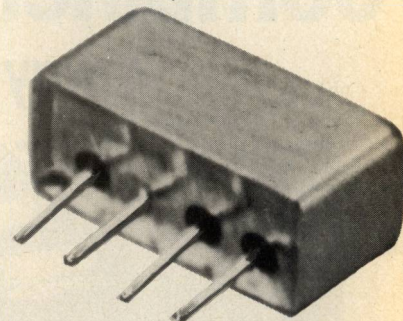


Fig. 8 One promising monolithic approach uses a zinc oxide overlay transducer to generate SAWs on a silicon substrate. Tap control electronics are integrated on the same chip. The SAWs are detected by piezoresistive MOSFET taps.

[Continued on page 146]

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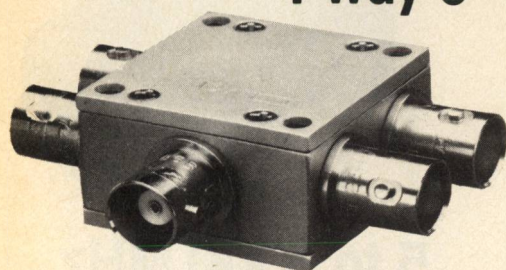
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PHASE UNBAL. (degrees)	1.0	4.0	
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[From page 145] FILTERS

On-chip switches select the desired phase by biasing one gate of each tap on while grounding the remaining gate.

The switches are controlled by on-chip logic circuitry. The input reference code can be read serially into groups of shift registers. Each shift register is connected through a latch to switch driver circuits. The latches are designed to hold the switches in the state corresponding to a particular code for correlation of an incoming signal while data for a new code is being read in.

Devices were fabricated by standard semiconductor processing techniques on three inch silicon wafers. The final chip size of 980 mils by 120 mils allowed 36 devices to be fabricated on a wafer. The chips were mounted on a thick film substrate for interconnections to package leads then housed in an all metal flat pack. Reproducibility and reliability of this process is good and offers potential for low cost in volume production.

Overall device performance was illustrated by correlation of various 31-bit maximal length sequences. Near theoretical peak to sidelobe ratios were obtained. Continuous operation was demonstrated through correlation of two alternating code sequences, each with a time duration equal to the total tap transit time of 3.1 microseconds. No degradation of correlator performance was observed under these conditions.

Monolithic GaAs

The development of GaAs technology for digital and microwave circuitry offers the promise of very simple and elegant programmable SAW transversal filters. Since GaAs is a weakly piezoelectric semiconductor, interdigital transducers can be fabricated directly on the GaAs surface when only moderate fractional bandwidths are required. This can greatly simplify fabrication procedures. There is in addition the potential for unique structures which utilize the interaction of the piezoelectrically active acoustic wave with the charge carriers in the material.

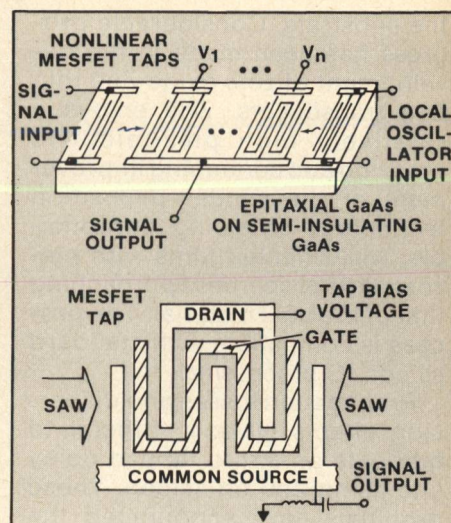


Fig. 9 A programmable filter on GaAs uses the nonlinear interaction of counter-propagating SAWs in FET mixer taps.

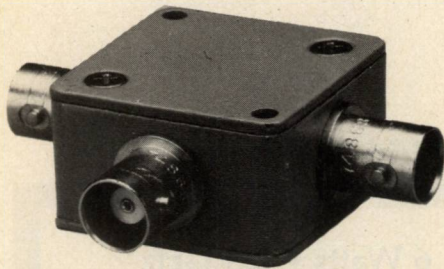
Take as an example the monolithic GaAs correlator fabricated by United Technologies Research Center.¹² This device uses the nonlinear interaction of two SAWs propagating in opposite directions through an array of MESFET taps as shown in Figure 9. One wave is generated by the signal while the other is generated by a CW local oscillator. The piezoelectrically generated fields associated with the SAWs interact directly with the fields in the FETs. The FET taps, shown schematically in Figure 9, were fabricated on an n-type epitaxial layer with a carrier concentration of $1 \times 10^{16} \text{ cm}^{-3}$ and a thickness of 1.3 microns. The source and drain contacts are ohmic and the floating gate electrode forms a Schottky barrier. When a DC bias voltage is applied between the source and drain, a quiescent current is established. The mixing of the counter propagating SAWs within the FETs produces components of the drain current at the sum and difference of the two SAW frequencies. The nonlinear signal power is proportional to the power carried by the SAWs and can be varied over at least 50 dB by varying the drain bias voltage. Reversing the polarity of the bias voltage shifts the phase of the output signal by 180°. Figure 10 is a photograph of the device showing 32 FET taps positioned between IDT's which generate the counterpropagating SAWs. The DC drain bias lines are

[Continued on page 148]

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INSERTION LOSS, dB	TYP. MAX.
one octave band edge	0.8 1.4
total range	1.5 2.3
DIRECTIVITY dB	TYP. MIN.
low range	30 20
mid range	27 20
upper range	22 10
IMPEDANCE	50 ohms

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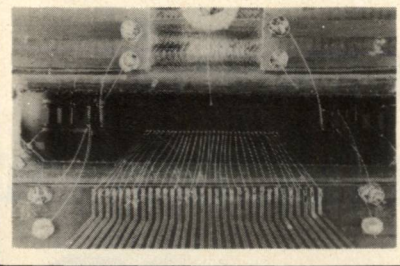
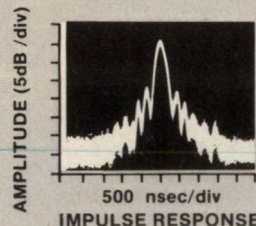


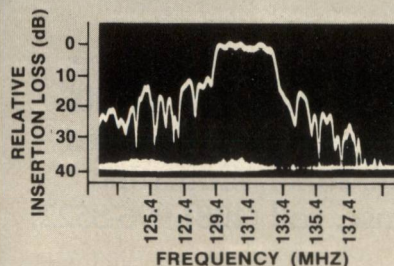
Fig. 10 A SAW correlator on GaAs with 32 programmable taps. (Photo courtesy of United Technologies Research Center).

visible at the bottom of photo with the RF output shown near the top. The center-to-center tap spacing corresponds to a sampling frequency of 10 MHz. The taps have a common source electrode for summing the nonlinear signals from the individual taps. The drains and gates are segmented allowing each tap to be independently programmed. The signal IDT has a center frequency of 127 MHz and a bandwidth of 9.5 MHz. The local oscillator transducer has a center frequency of 110 MHz and a 4 MHz bandwidth. An external bandpass filter selects the output signal at the difference frequency.

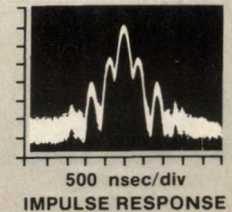
When used as a matched filter for biphase signals, the coded waveform is correlated with the code programmed on the taps.



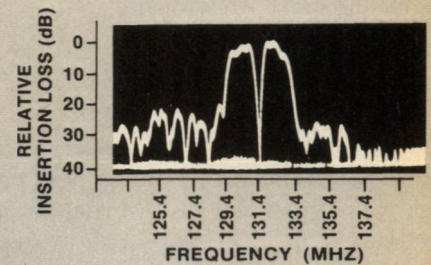
11 a. Impulse response of GaAs filter programmed for a $\sin x$ response.



11b. The resulting 3.8 MHz bandpass filter response



12a. Modified impulse response used to produce a null in the filter passband.



12b. Frequency response.

Near ideal sidelobe response has been obtained for the 13 chip barker code and for 31 chip pseudo-random maximum length codes. The device exhibits a dynamic range of greater than 65 dB and spurious signal suppression of at least 50 dB.

Operation of this device as a programmable frequency filter is demonstrated in Figures 11 and 12.¹³ Figure 11b shows the frequency response of a 3.8 MHz bandpass filter realized by programming the taps to obtain the $\sin x$ impulse response of Figure 11a.

The impulse response of Figure 12a was calculated to place a null in the passband as shown in Figure 12b. The 40 dB null would be very effective in suppressing narrowband interference and demonstrates the accuracy that can be achieved with this device.

Conclusions

Three general approaches to obtaining compact, rugged programmable SAW transversal filters have been described. The filters considered here fall into various stages of development, from R&D prototype to pilot production models. For some, the level of

[Continued on page 150]

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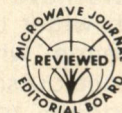
[From page 148] **FILTERS**

integration is just sufficient to show programmability and compatibility with peripheral electronics, while others include relatively sophisticated tap control electronics.

In all cases, the potential for large time bandwidth product, wide dynamic range and low spurious response has been demonstrated. With continued development, this SAW technology will provide key components to meet evolving systems requirements for adaptive frequency filtering and asynchronous matched filtering.

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Low Noise GaAs FET Dual Channel Front End

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National Astronomy &
Ionosphere Center
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Ithaca, NY

Low noise receiver systems are degraded by the noise contribution of the transmission line which connects the receiver to its antenna. Receiver performance and behavior are affected to some extent, by antenna impedance variations across the frequency band of amplification of the RF amplifier. This paper describes the L-band front-end designed to minimize effects which are external to the amplifier. The design results in minimal receiver equivalent noise temperature at the waveguide flange which connects directly to the antenna.

The design of low noise microwave amplifiers using Gallium Arsenide Field Effect Transistors (GaAs FETs) is well understood. Several authors have described state-of-the-art development of these devices.¹⁻² When operated at cryogenic temperatures GaAs FET amplifiers provide competitive performance to cooled parametric amplifiers without the complexity and higher costs associated with paramps.

The Mitsubishi type MGF-1412 GaAs FET was selected for the design of the L-band amplifier although several other low noise GaAs FET devices are currently available.

The input impedance of a GaAs FET is predominated by capacitive reactance. The gate to source connection exhibits essentially the same characteristics as that of a capacitor, consequently the first step toward achieving an input match is to tune out the capacitive reactance. At 1420 MHz this is accomplished by a series inductor of about 8 nanohenries.

The desired 50 ohm real part of the input impedance is achieved by adding an inductor in the source lead of the FET. This applies a negative feedback to the transistor hence increasing the real component of the impedance. This procedure has been described.¹⁻³

A high pass filter is included in the input circuit, and bias voltages are supplied via appropriate filters, shown in the schematic of Figure 1.

Similar matching is used to tune the output to 1420 MHz after which the return loss is 15 dB or better for both output and input ports.

Recent designers of cooled microwave GaAs FET amplifiers have achieved excellent impedance matching by means of source inductance feedback. L-band amplifiers using this technique have been described by Nevin and Wong.³ A similar L-band amplifier cooled to cryogenic

temperatures is described by Williams, Weinreb and Lum.⁴ Cooled GaAs FET amplifiers of this general design have been used at the NAIC Radio Telescope near Arecibo, Puerto Rico with limited success because of certain inherent characteristics of the antenna.

The Arecibo 1000' spherical reflector (see *World's Largest Radar*, this issue) is illuminated by various single and dual polarized feeds. A line feed exhibits a relatively narrow bandwidth, for example the 21 cm dual polarized feed has a half powerpoint bandwidth of about 70 MHz. The impedance match is excellent ($VSWR \leq 1.1:1$) within the bandwidth of the antenna; but, like a narrow band filter the impedance match degrades rapidly and severely outside the bandwidth of the antenna.

This characteristic of the Arecibo telescope antenna has caused unstable operation of some of the

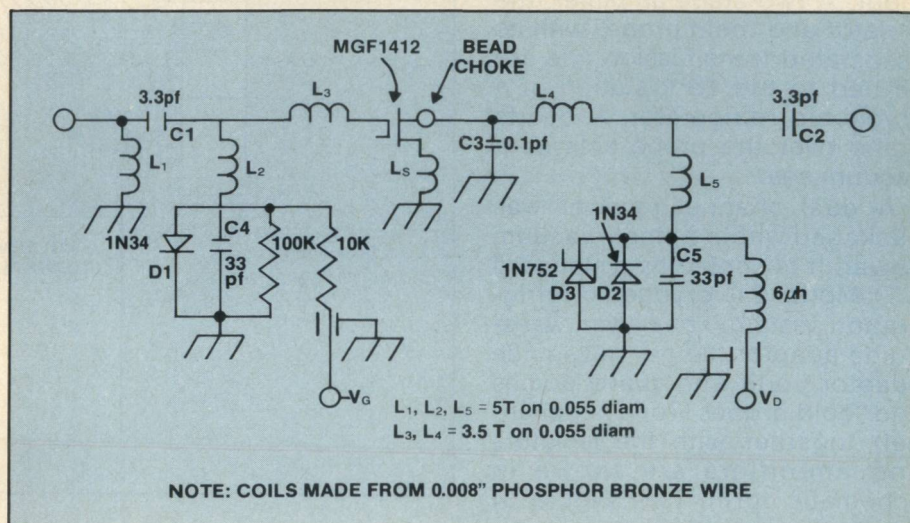


Fig. 1 GaAs FET RF amplifier.

GaAs FET amplifiers tested. The amplifiers have substantial gain at much wider bandwidths than the antenna, while the antenna impedance outside its bandwidth is unknown to the amplifier designer, and, even if known it would not usually be practical to take it into account in the design.

Thus, a receiver intended for use on the Arecibo telescope feeds logically should incorporate a good isolator to prevent antenna impedance variations from affecting the amplifier behavior. Unfortunately isolators typically introduce 0.2 db to 0.4 db in-band insertion loss. When the isolator is operated at room temperature this represents an unacceptable noise contribution of 14° k or more by the isolator. From previous experience it was evident that isolators could be developed to operate at cryogenic temperatures. When operated at 16° k, the low 0.2 db insertion loss would contribute only about 1° k to the receiver temperature.

Another source of noise in a conventional amplifier is the transmission line loss within the cooled dewar and the coax to waveguide adaptor necessary to connect to the antenna. A technology to minimize transmission line losses was borrowed from experience gained in the S band maser development.⁵ An inner conductor of a coax line is extended into a WR650 waveguide adaptor port. (See Figure 2) The "cold probe" is designed to be matched for a low VSWR at its connection to the cooled isolator input. A two-stage amplifier, the isolator and "cold probe" with its associated transmission line are all tied to the 16° k station of a cryogenic refrigerator. A quartz dome over the probe acts as a vacuum seal.

A dual channel receiver was packaged within a single vacuum dewar. It is cooled by a standard CTI Model 21⁶ cryogenic refrigeration system. The dewar, waveguide adaptor flange, waveguide adaptor ports with quartz domes and "cold probes" (one per channel) together with the isolators and amplifiers are shown in schematic form. (see Figure 3) Precautions were taken to minimize heat loss and to assure long

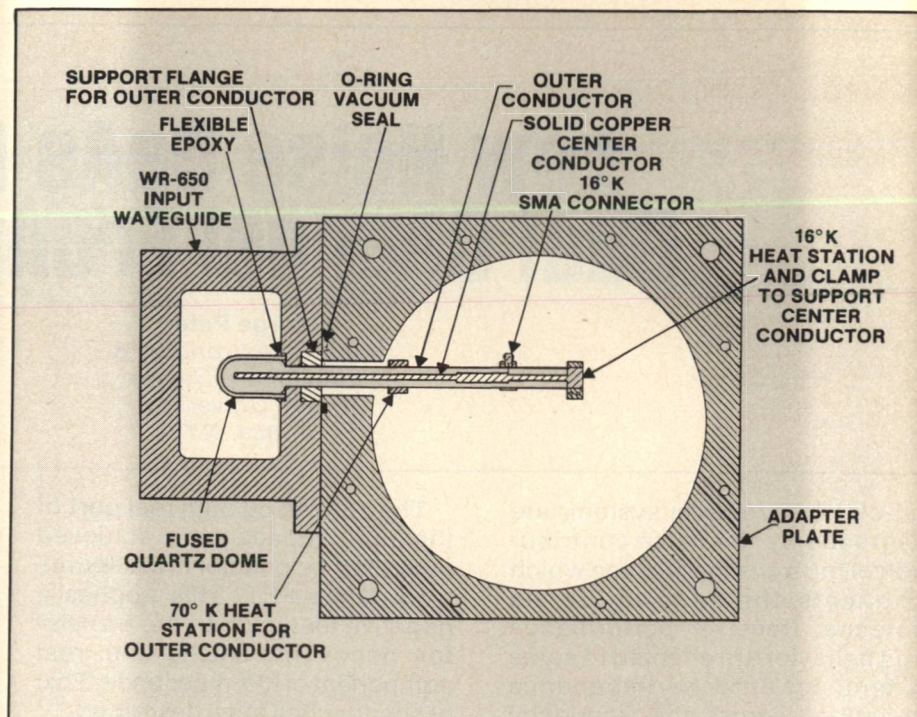


Fig. 2 Single channel of cold probe assembly.

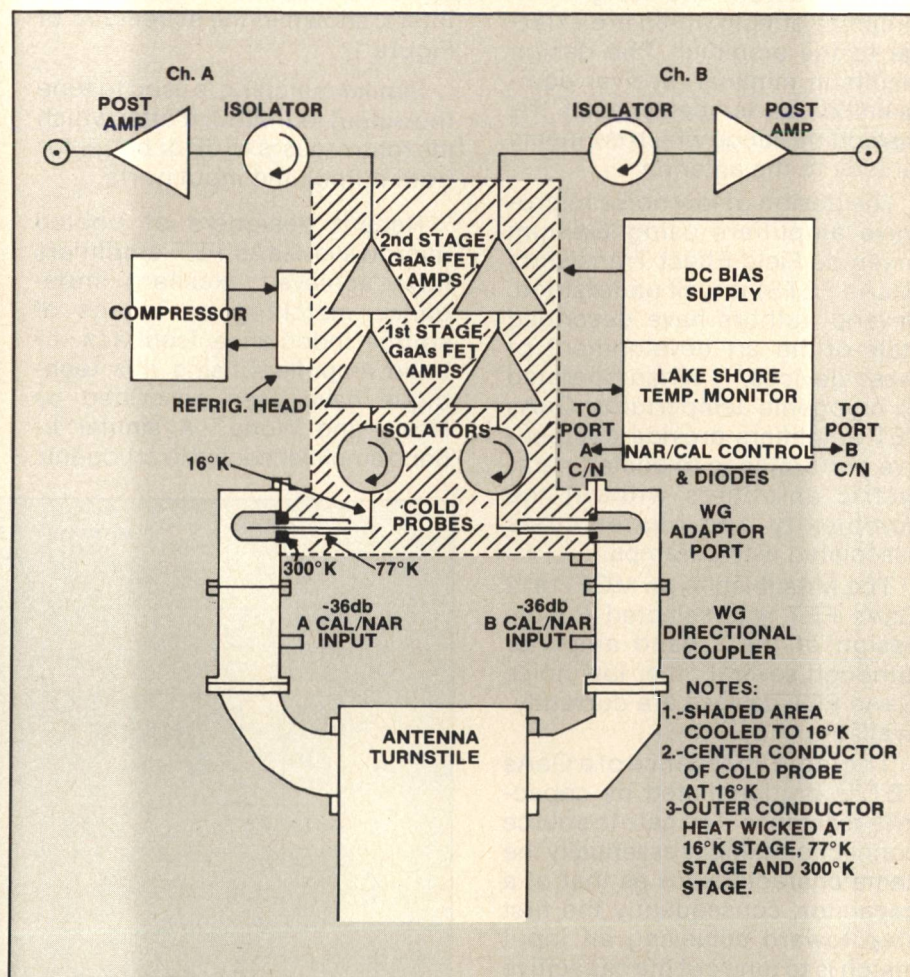
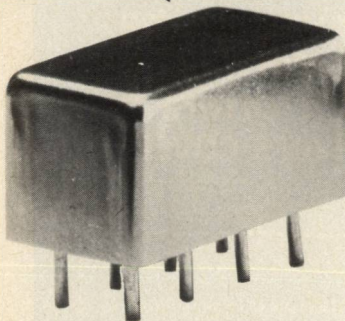


Fig. 3 Block diagram of RF dual channel GaAs FET cooled front end.

[Continued on page 156]

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IF .05 - 500

CONVERSION LOSS, dB

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Total range

ISOLATION, dB

.05-.5

.5-1000

1000-2000

	TYP.	MAX.
LO-RF	6.0	7.5
LO-IF	7.0	9.0
	TYP.	MIN.
LO-RF	25	20
LO-IF	25	20
LO-RF	40	30
LO-IF	40	30
LO-RF	30	20
LO-IF	25	15

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[From page 154] FET

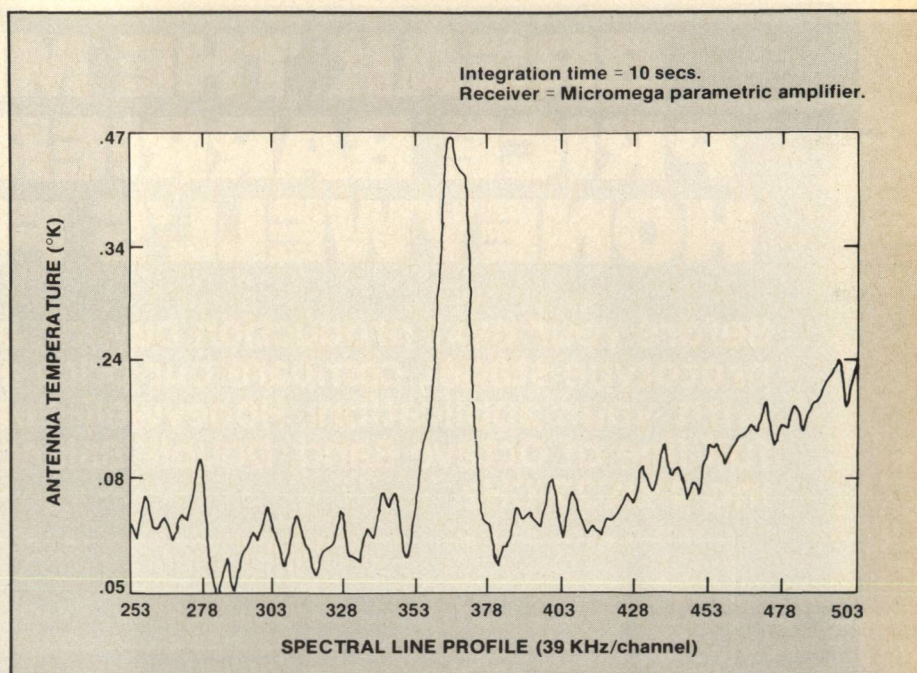


Fig. 4a Signal from source UGC 3376 as compared to background noise.

term reliable operation. A separate package attached to the dewar provides GaAsFET bias supplies, metering and calibration signals. The total mechanism is designed to mount directly above the antenna turnstile thus keeping the waveguide lengths to each port of the dual polarized feed to a minimum (<2').

The system was installed at the Arecibo facility in early December 1980 and has been in operation

since. It provides a stable low noise receiver system with excellent baseline. (See comparison of original parametric amplifier results with those of the new GaAs FET amplifier system.) See Figure 4, a and b)

Receiver equivalent noise temperature at the waveguide flange is measured to be <15°K across the band from 1300 to 1500 MHz. System temperature (including antenna) is calculated to be 40°K,

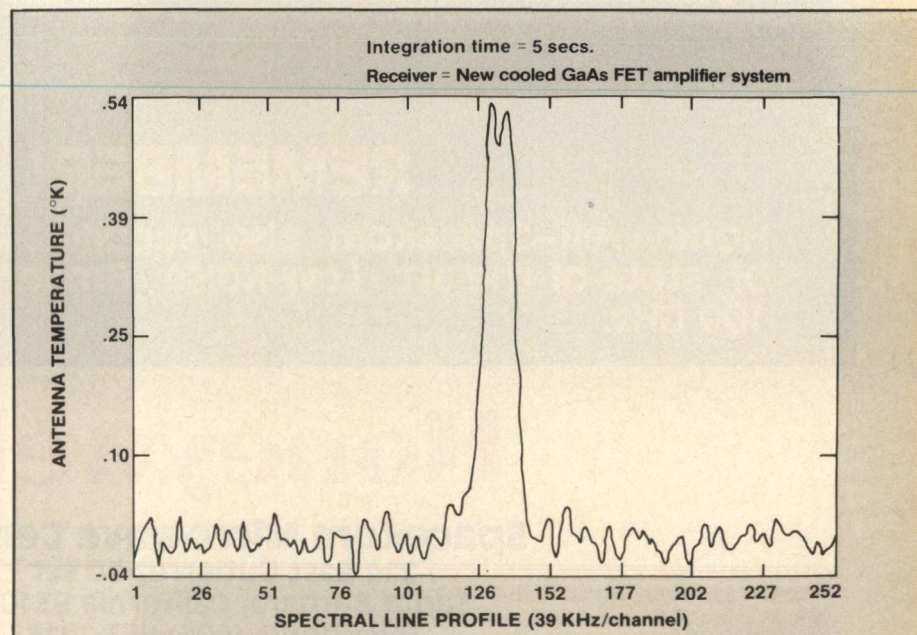
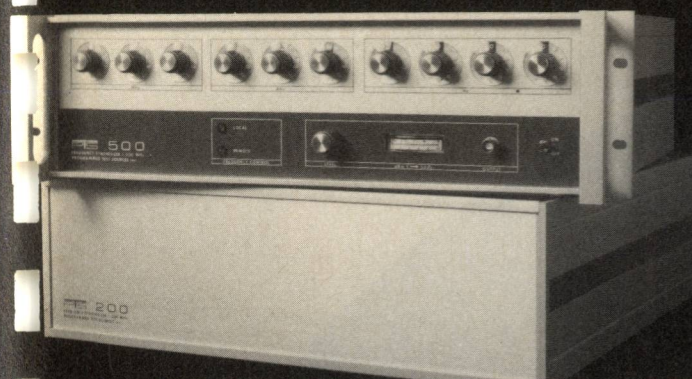


Fig. 4b Signal from source UGC 3376 as compared to background noise.

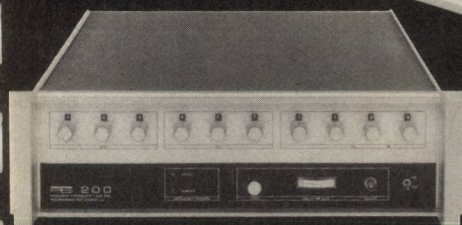
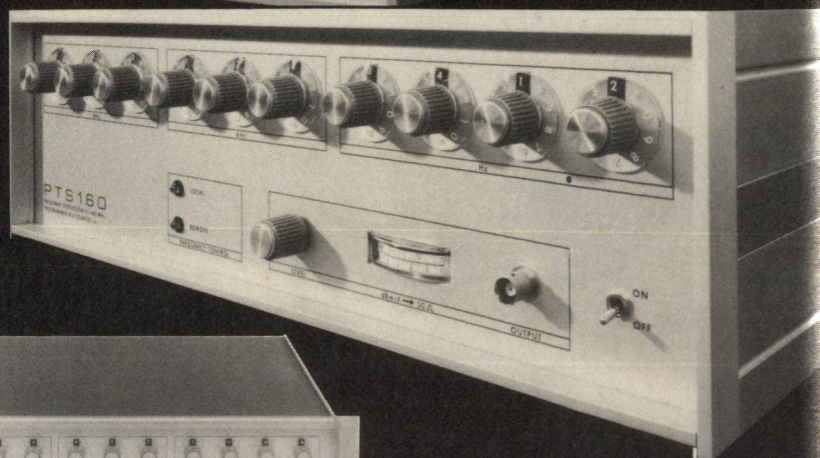
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[From page 156] **FET**

in which noise contribution of the antenna has been estimated to be 20°K and sky noise contribution is about 5°K.

In summary three technologies have been incorporated to achieve a stable low noise, L-band front-end. The development of GaAs FET amplifiers and the design of optimum match for minimal noise contribution have been borrowed from other designers. Commercial manufacturers were encouraged to develop L-band isolators to operate at cryogenic temperatures. The development of a "cold probe" to operate at L-band was the final technology which, when incorporated with the cooled amplifiers and isolators, resulted in a quality low noise dual channel receiver system.

Acknowledgements

The National Astronomy and Ionosphere Center (NAIC) is operated by Cornell University under contract with the National Science Foundation. Dr. Donald Campbell, is the Acting Director of NAIC.

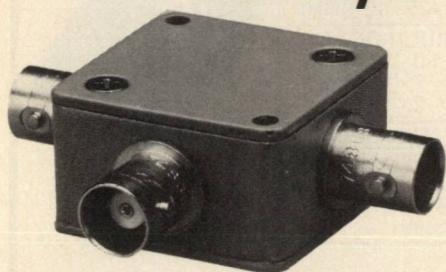
My colleague Lynn Baker's support, advice and ideas helped to develop the receiver system. Credit is due the staff in the development process: Linda English, William Genter, Kurt Kabelac, Pierson Mosher and David Van Winkle. Engineers Edgar Castro and Robert Zimmerman of the Arecibo facility acted as coordinators to ensure compatibility with existing equipment of the Observatory.

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6. Model 21 Cryodyne, CTI-Cryogenics, Inc. Waltham, MA

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INSERTION LOSS,		
above 3 dB	TYP.	MAX.
10-100 MHz	0.25	0.6
100-750 MHz	0.5	1.0
750-1500 MHz	0.8	1.5
ISOLATION, dB	25	
AMPLITUDE UNBAL., dB	0.2	0.5
PHASE UNBAL., (degrees)	5	10
IMPEDANCE	50 ohms	

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Application Note

Biphase and Quadriphase Digital Modulators

Robert Moody

Anzac Div., Adams Russell
Burlington, MA

Introduction

Digital modulators provide the basic building blocks for bridging the boundary between the quantized video world of the data user or digital designer and the RF oriented discipline of the communicator or communications system designer. The modulators accept a digital data stream and an RF carrier input and produce a phase modulated output. The properties of the output signal are dependent upon those of the inputs, the variable source is the data waveform.

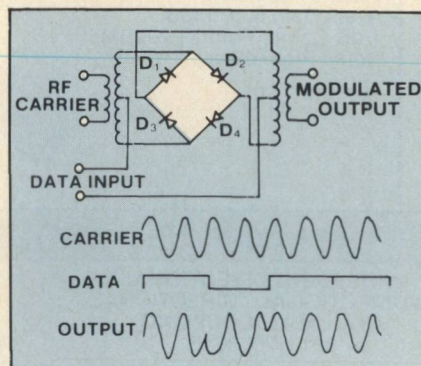


Fig. 1 Biphase modulator.

Biphase Modulators

Biphase Modulators utilize a circuit similar to that of doubly balanced mixers (Figure 1) as the basic modulating device. A carrier signal injected at the RF input will appear at the RF output reduced slightly in amplitude due

to inherent diode and transformer losses in one of two possible phase states with the modulating signal and the carrier suppressed. In the 0° phase state, control current is supplied to the diode ring to turn on a pair of diodes (either D1/D4 or D2/D3) which directly connect the input and output transformers. In the 180° phase state, current of opposite polarity switches on the other pair of diodes causing a 180° phase inversion relative to the 0° state. If the modulating bit rate is near the carrier frequency, a waveform of the kind shown in Figure 1 might be obtained. This waveform can be interpreted as a biphase signal or as a double sideband suppressed carrier modulation since in this case they are equivalent.

Quadriphase Modulators

The basic modulating elements in a quadriphase or QPSK modulator are a pair of matched biphase modulators. The carrier signal applied to the RF input first passes through a quadrature hybrid. The two resulting carrier signals now having a 90° phase relation then pass through the matched pair of biphase modulators, and finally the resulting signals are added vectorially in an output power combiner as shown in Figure 2. In this case, two control bits at a time are used, each controlling the phase of one of the orthogon-

al components of the output signal. It is assumed that any serial to parallel conversion required to generate the simultaneous modulating bits will be performed with external logic.

Four phase states are possible with this configuration. The reference or 0° state occurs when both biphas modulators are in the straight through or 0° state. In this case, the two signals from the quadrature hybrid at 0° and -90° relative phase appear unshifted at the output power combiner and add vectorially to give a resultant insertion phase shift of -45° . The other three states occur when the biphas modulators are in the 0° and 180° states, 180° and 180° states, and 180° and 0° states, giving respective insertion phase shifts of $+45^\circ$, -135° and $+135^\circ$.

The circuit element that generally limits the usable range of carrier frequency in a QPSK modulator is the quadrature hybrid.

Digital Modulator Functions

Both biphas and QPSK modulators transform a stream of digital "1" and "0" bits into phase shift keyed modulation on an RF carrier. While the RF parameters of a biphas or QPSK modulator can be determined by static measurements, it is important to consider the actual operation with digital signals applied. The bit pattern of the data waveform will tend to be random with the exception of any overhead bits used for framing or synchronization.

Figure 3 shows the envelope of the baseband spectrum and the output from the phase modulators. As indicated, the quadriphase modulators will produce the same spectrum as the biphas unit with the appropriate 2 to 1 scale change for frequency. The width of the main lobe envelope of the ((sin

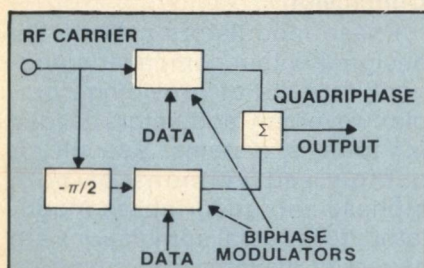


Fig. 2 Quadriphase modulator.

$x)/x)^2$ distribution is determined by the bit rate of the modulating sequence and the line spacing under the envelope is determined by the repetition rate of the sequence.

Most of the energy in the spectra is contained in the main lobe, only about 10% is contained in the sidelobes; and the spectral bandwidth is directly dependent upon the data bit rate.

The data rate often is an indication of the type of logic which will be supplying the modulating signal and the type of driver used in the modulator.

The two major types of drivers used are TTL and ECL. The TTL logic employing Schottky type elements will function at bit rates up to near 100 Mbps, while emitter coupled logic can extend operation into the 500 Mbps region.

Modulator Characteristics

Spurious Spectrum Components: The previous section has described the general character of the input data stream and the resulting spectra observable at the modulator output. The data user or digital designer is often less interested in the spectra shown in Figure 3 than in the measurement of the quality of the recovered data sequence at the demodulator output. However, there are spectral properties which may affect system operation which are not readily deduced by studying the recovered data stream. A modulator output spectrum contains a variety of spurious signals which might be treated by a demodulating system as interference. The component most often specified is the suppressed carrier. The carrier component is in band and can be reduced by

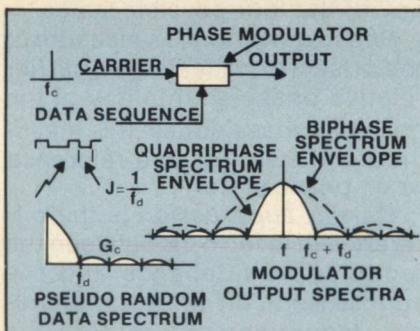
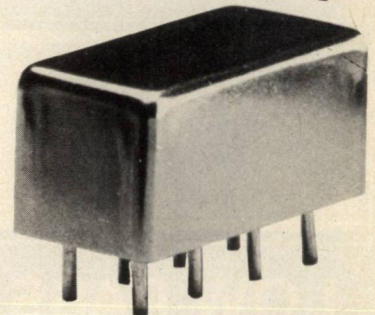


Fig. 3 Data and modulator spectra.

[Continued on page 162]

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100-1000 MHz	0.7	1.2
ISOLATION, dB	25dB	TYP.
AMPLITUDE UNBAL.	0.2	TYP.
PHASE UNBAL.	2°	TYP.
IMPEDANCE	50 ohms.	

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[From page 161] MODULATORS

improving modulator balance. Levels of 30 dB or more carrier suppression are commonly achieved in well designed biphase and QPSK modulators.

Insertion loss of a biphase or QPSK modulator is measured statically in either of the possible phase states and represents the worst case loss for any state.

Biphase modulators theoretically would have no loss when measured in this manner. However, finite losses do occur because of transformer and diode dissipation. Typical losses of 3-3.5 dB are to be expected. QPSK modulators have a theoretical insertion loss of 3 dB to the quadrature phase relation of the two signals fed to the output power combiner. Additional dissipative losses in the biphase modulator, quadrature hybrid and power combiner increase this loss. Typical losses of 5.5-6.5 dB are to be expected.

VSWR is a measure of the impedance mismatch at the RF input and output ports of the modulator. It is of importance because the presence of imperfect source and load impedances in real system application combined with mismatches at the input and output of the modulator may introduce phase errors due to multiple reflections. For this reason, the VSWR of modulators should be tightly controlled, typical values are 1.3:1.

Amplitude balance is a measure of the variation in insertion loss between the possible phase states of the modulator. For biphase modulators, it is simply the variation in insertion loss between the 0° and 180° states, while for QPSK modulators it is a measure of the worst case of variation between any of the four possible states.

Phase deviation is a measure of the offset from the desired carrier relative phase shift in any of the possible phase states. It is measured with respect to a reference or 0° phase state.

Carrier suppression is the difference in level in dB between the modulation sidebands and the suppressed carrier. It is dependent upon symmetry in both the modulator circuitry and in the

modulating waveform. Carrier suppression is typically 35 dB or greater in well designed biphase and QPSK modulators.

Modulator Logic Interface

Biphase and QPSK modulators are generally available with or without integral drivers. The modulators containing ECL drivers simplify the designer's task by standardizing the logic interface while still allowing the high data rates typical of ECL logic. The interface to the control ports of these devices should follow standard ECL practices. Typical control voltages are -0.9 V for a high state and -1.75 typically for a low state. (See Figure 4).

Modulators without integral drivers typically require an external driver capable of delivering ± 10 ma. Care should be taken in design of the external driver to maintain symmetry of the positive and negative half cycles of the driving waveform and RF interfacing practices appropriate to the data rate should be used in connecting the driver to the modulator.

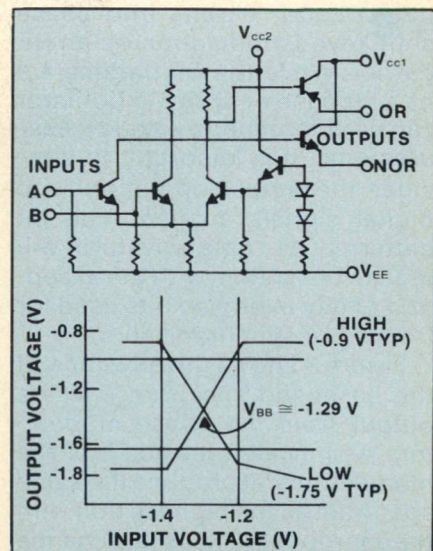


Fig. 4 Typical ECL and transfer characteristics.

Conclusion

Phase modulators provide the designer with a compact component capable of providing complex waveform modulators needed for today's systems. Specific or customized versions of both biphase and quadriphase modulators for special applications are generally available from suppliers of the basic components. ■