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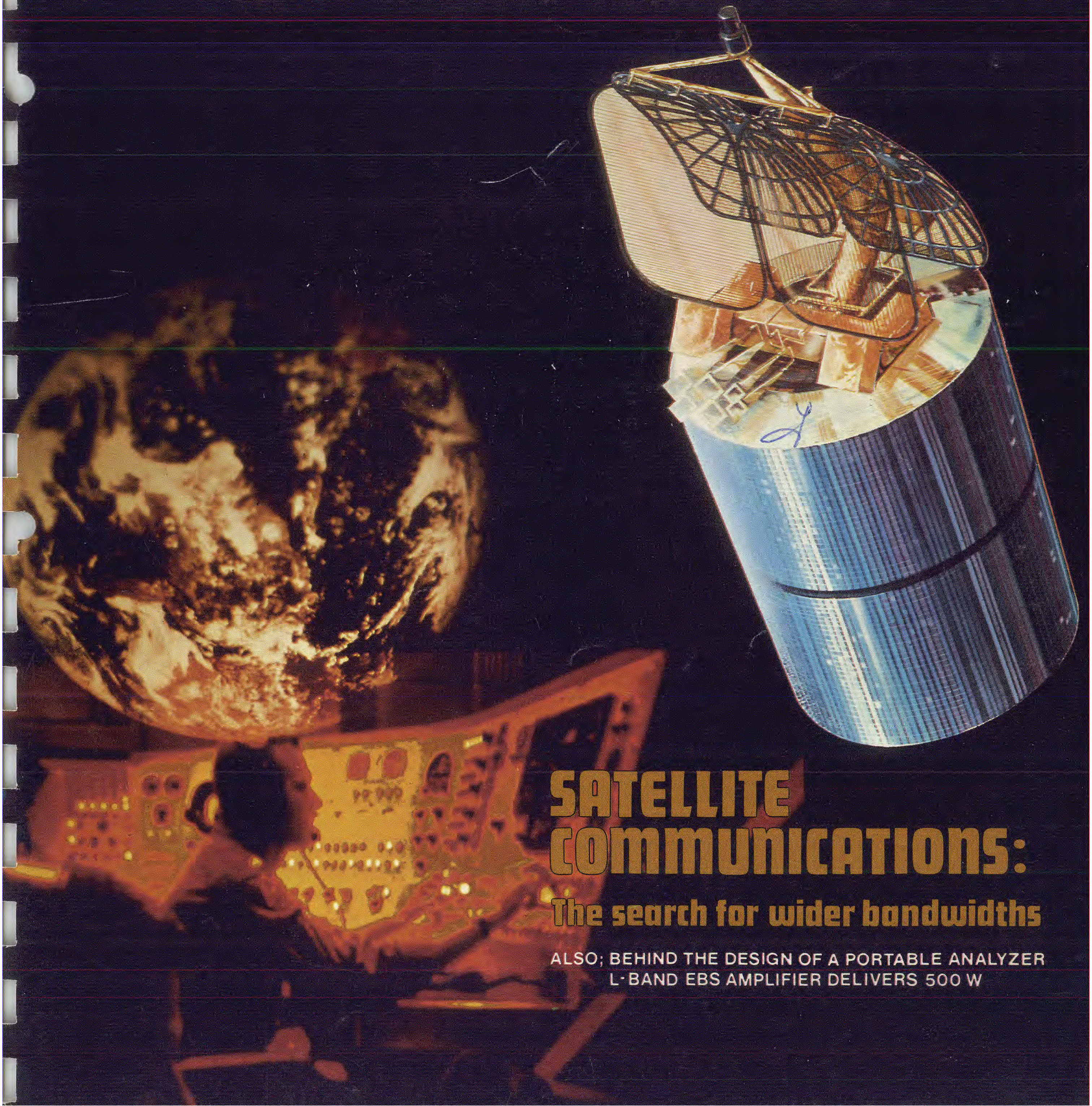

SATELLITE COMMUNICATIONS:

The search for wider bandwidths

ALSO; BEHIND THE DESIGN OF A PORTABLE ANALYZER
L-BAND EBS AMPLIFIER DELIVERS 500 W

MICROWAVES

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About the cover: Comsat General's domestic satellite for 1976, designed and built by Hughes Aircraft, is squeezing out additional channel capacity at the 4 and 6 GHz bands by using polarization diversity. It will also be evaluating the millimeter bands for possible future use. Photo composition by Robert Meehan.

coming next month: Semiconductors

HYBRID COUPLED AMPS: CAN THEY WEATHER A MISMATCH? Richard LaRosa of Hazeltine in Greenlawn, NY, examines the effects of an output mismatch on a pair of hybrid-coupled power transistors. The analysis reveals that one transistor is in jeopardy, even with a relatively minor mismatch. Surprisingly, this problem cannot be detected by monitoring the terminated hybrid output port.

MODELING TRANSISTORS FOR IMPROVED PERFORMANCE. Jim Curtis of PHI describes a practical method of modeling power transistors. The lumped-equivalent circuits derived are based on transistor loading. Large signal conditions apply.

Also: The 1974 Annual Index

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L-Band EBS amplifier belts out 500 W

Stacy V. Bearse
Associate Editor

Specialists in vacuum tube and solid-state technologies have teamed up to push a high-power electron bombarded semiconductor amplifier into the gigahertz region. With potential TACAN and IFF applications in mind, researchers at Watkins-Johnson in Palo Alto, CA have designed a developmental EBS amp that produces more than 500 watts in L-band.

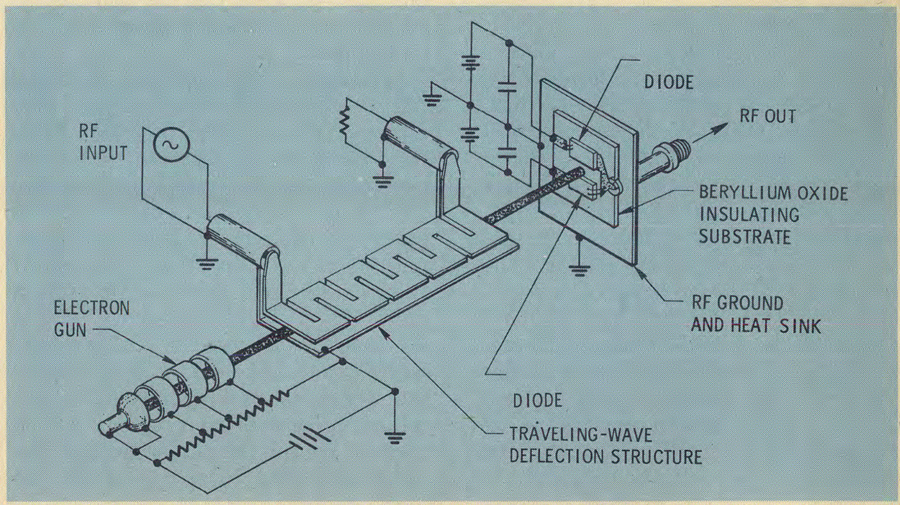
"The maximum output power obtained at 1.5 GHz is 520 W at 0.1% duty and 250 W at 1% duty. At the 500 W level, the saturated gain is 23 dB and the small signal gain is 29 dB. The 3 dB bandwidth is 106 MHz or about 6.8%," details James A. Long., a member of the technical staff of W-J's power tube department.

EBS amplifiers are hybrid devices that combine an electron-gun assembly, traveling-wave structure, semiconductor targets and an output coupling network within a single vacuum envelope (Fig. 1). "We're developing them because they offer significant advantages over triodes and solid-state devices in terms of power-bandwidth product and gain linearity," the W-J researcher notes.

W-J's EBS device, announced at the 1974 International Electron Devices Meeting, amplifies the current flowing though a pair of silicon diodes by belting them with a 20 kV electron beam. Before striking the semiconductor targets, the electron beam is modulated by passing it through a broadband meanderline structure. Here, an rf input deflects the beam and produces linear modulation.

Four years ago, a 400 W, 100 MHz EBS amplifier drew a lot of attention ("Hybrid tube-semiconductor device developed as high power amp", MicroWaves, February, 1971). The ability to get better performance at more than 10 times the frequency is largely due to the development of an improved output tuning circuit, Long indicates.

"The new amplifier uses a novel diode interconnection circuit which connects two diodes in rf series while they are biased in dc parallel (Fig. 2). Operation of the diodes in rf series is an advantage since the reactance level of a typical high-power diode is only a few ohms at microwave frequencies. By connecting two or more in series, it is possible to increase the load impedance into which the circuit will operate efficiently," he says.



1. EBS devices use a traveling-wave structure to modulate an electron beam. Bombarding the semiconductors with the modulated beam causes a current multiplication of approximately 2000:1.

Both diodes are isolated from ground and from each other by quarter-wave chokes, while two-radial line resonators and a coupling capacitor provide an impedance match to the 50-ohm output line.

"The sectorial, radial resonators were designed by a combination of computer-aided and empirical procedures. The outer radius and coupling capacitance were determined by trial and error in a computer program to achieve the center frequency of 1.5 GHz, and the center frequency resistance of 10 to 15 ohms," Long explains.

Laboratory tests showed an unloaded resonator Q of 800.

Production problems cited

What remains to be done before this developmental design can be translated into practical hardware?

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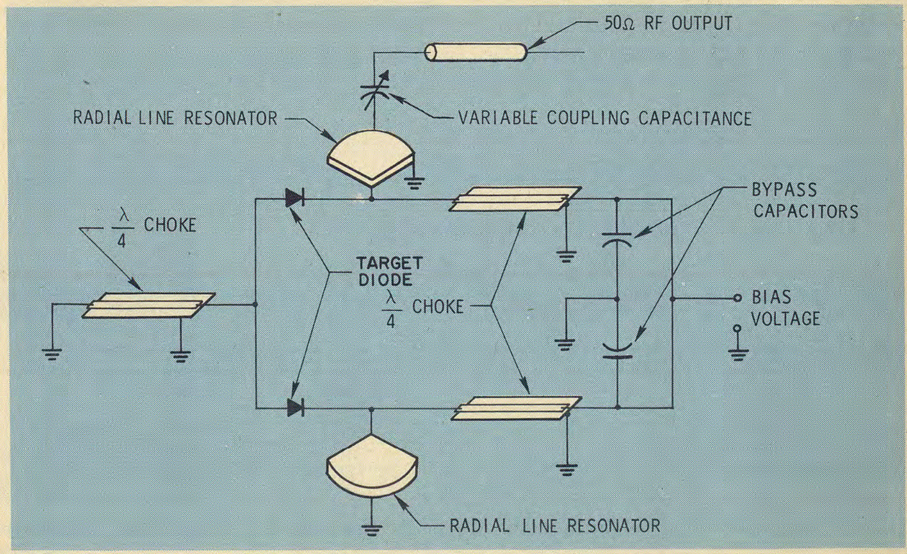

2. The development of an effective target tuning circuit was the key to high-power, gigahertz performance.

NY crime fighters now using laser beams for fingerprint identification

Howard Bierman
Editor/Publisher

Latent fingerprints left at the scene of a crime in New York City will now be rapidly scanned, coded and matched by a semi-automated system involving a laser beam technique. New York Police Commissioner, Michael J. Codd, recently announced this latent fingerprint system, designed by the McDonnell Douglas Electronics Co., that combines microfilmed suspect prints with a laser beam arrangement to provide data processing of keypunch input with optical scanning of fingerprint images at a speed of four comparisons per second.

"Prior to this system, single fingerprints had to be manually scanned, coded and matched against N. Y. City's Police Department's active file of over 130,000 major crime offenders," explained Dr. Marvin Berkowitz, Director of Applied Technology for the Police Department.

Matched filter correlation key to system

Matched filter correlation, using coherent optical processing techniques, is the heart of the latent fingerprint identification system. The optics approach offers several advantages over an all-electronic system: (1) there is no need to examine each fingerprint and describe the video wave-form prior to processing, (2) the optical approach permits simultaneous comparison of multiple fingerprints in a library file with the latent fingerprint and (3) a complex mathematical process can be implemented with a relatively simple set of equipment.

Basically, the optical matched filter correlation technique produces a measurable physical quality (light intensity) which is proportional to the similarity of the fingerprints being compared. In other words, if a latent fingerprint were to match a fingerprint in the library file, a maximum amount of light intensity would be produced when both prints were super-imposed. A non-matching fingerprint, compared to the latent fingerprint taken at a crime, would block some of the light and less intensity would be measured.

The Fourier transforming property of lenses forms the basis of the design of the electro-optical system. If a photographer transparency of a two-dimensional pattern (such as a fingerprint) is placed in the front focal plane of a lens and illuminated by a collimated right beam (laser), the light pattern which appears at the back focal plane of the lens is a Fourier transform of the input pattern. The inverse Fourier transform of the product of the two Fourier transforms (one from the latent fingerprint at the crime, the second from the library file), is a direct measure of the similarity and is the correlation function. By this technique, the electro-optic system can sort out of a library file those fingerprints that are most similar to the one being checked.

Microfilm transparencies of latent and library fingerprints are used to introduce them into the system for comparison. In the comparison process, the inverse Fourier transform of the product of the Fourier transforms of the fingerprints being compared is measured and probable matches are determined. This is accomplished by:

- Recording an interferogram of the Fourier transform of the latent print on photographic film, which provides a latent filter.
- Illuminating this filter with the Fourier transform of the library fingerprints.
- Focusing the light emerging from the filter.
- Measuring the peak intensity of the focused light to determine the correlation or degree of match between the fingerprints.
- Repeating the comparison process with each suspect contained in the library subfile and selecting those which exhibit the highest correlation as the most probable matches.

The Fourier transform of a fingerprint transparency is produced by an interferometric technique. A beam of light from a laser is split into two beams which are then separately filtered and collimated, Fig. 1. The lower beam (reference) is redirected by a prism to intersect the top beam (signal) at plane B-B. A Fourier lens is positioned in the top beam

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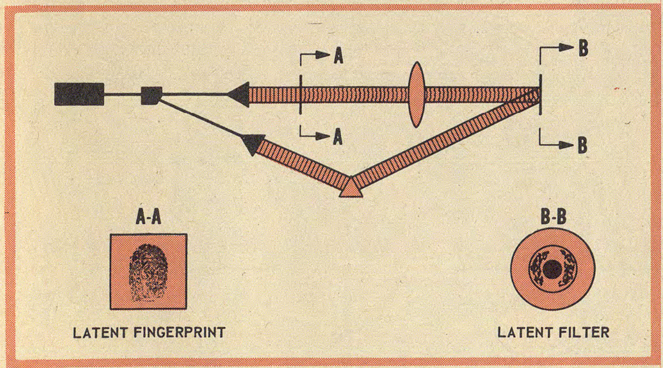


1. An interferometric technique is used to develop a Fourier transform at the back focal plane (B-B) from a fingerprint transparency placed at the front focal plane of the lens (A-A).

2. The McDonnell Douglas latent fingerprint system costs N. Y.'s Police Dept. $300,000 largely covered by a federal grant from the Law Enforcement Assistance Administration.

NY crime fighters now using laser beams for fingerprint identification (cont.)

(continued from p. 10)

so that its back focus corresponds to the beam intersection (B-B). A microfilm of the latent fingerprint is placed in the top beam at the front focal plane (A-A) of the lens, causing its Fourier transform to appear at the back focal plane. Unexposed film is placed at B-B, and the interferogram is recorded. The bottom (or reference) beam is required to enable both phase and amplitude of the signal beam to be captured on film.

In the comparator section, a beam of light from a laser is expanded and passed through a microfilm aperture card containing the library suspect fingerprint to be compared with the latent fingerprint. The library aperture cards are sequentially placed in the front focal plane of a simple lens at a rate four per second by a card handler. The latent filter of the unknown fingerprint is placed at the back focal plane of the lens and resultant output of the filter is the product of the library and latent fingerprint Fourier transforms. The intensity of the resultant point of light is the desired correlation value of the fingerprint.

During the evaluation of each library aperture card, the average diffraction efficiency of the entire card is measured. This information is used by the video processing electronics and decision logic to evaluate that card for probably match. Also, the latent fingerprint filter is rotated during the evaluation process to compensate for the angular orientation difference between the library and latent fingerprints. The correlation light energy is a function of proper orientation between the fingerprints being compared. • •

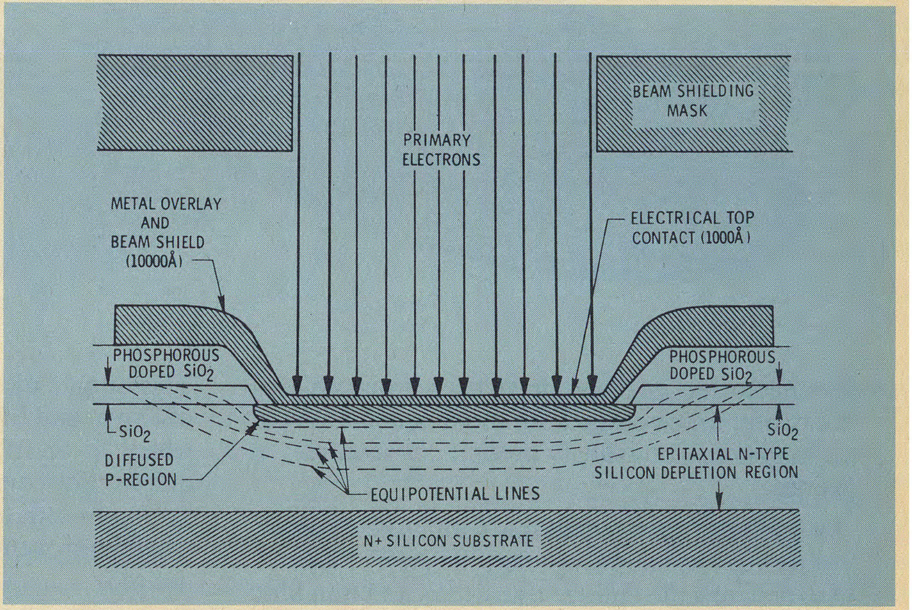
L-Band EBS amplifier belts out 500 W (cont.)

(continued from p. 9)

"We have some simplification that has to be done in the mechanical design, but I think that the principal thing that we have to do is develop a reliable target diode," Long comments.

The semiconductor targets have an epitaxial, n-type, silicon depletion region on top of an n+ silicon substrate (Fig. 3). A diffused P-region is covered by a thin aluminum top contact and is surrounded by a thick aluminum overlay which serves as a beam shield and an electrical contact. The metal overlay is separated from the depletion region by a layer of thermally-grown silicon dioxide and a layer of deposited phosphorous-doped glass to provide passivation and increase the reverse voltage breakdown. The depletion region is approximately 17 microns thick and has a resistivity of 10 ohm-centimeters.

In operation, a copper beamshield is placed in front of the diodes allowing electrons to strike the active area without bombarding the peripheral silicon and metal overlay.



3. The semiconductor targets have an active area of 2.5 x 1.0 mm. A beam shielding mask protects the peripheral silicon and metal overlays.

Although the diode targets for the developmental tube were fabricated by the Signetics Corporation, Long says that W-J is currently developing an in-house production facility. ••

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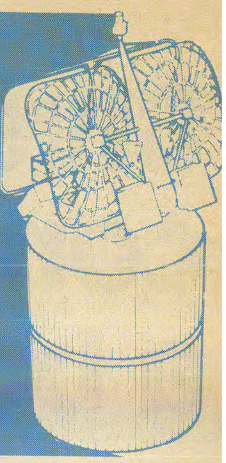

Satellite communication: The search for more bandwidth

Richard T. Davis
Managing Editor

How to get more channel capacity for the dollar appears to be the major thrust for the next group of domestic satellites planned for launch in 1976. Typical of this thinking are the vehicles being developed by Comsat General for AT&T and by RCA Globecom. Both satellites will be using polarization diversity to effectively double their channel capacity while still only using the 500 MHz bandwidths available at the 4 and 6 GHz satellite bands. By using orthogonal linear polarization to achieve frequency reuse, the satellites will be able to carry 24 transponders, each providing 1200 voice grade circuits or twice the capacity of Western Union's Westar and the Canadians' service via Anik.

Polarization diversity isn't new to communications—at least terrestrial communications. "It's long been used on AT&T's TD and TH microwave radio links but without spectrum overlap," says Lance Froome, Member Technical Staff, Satellite Systems Engineering Department, Bell Labs, Holmdel, NJ. "But for satellite communications, using overlapping channels, the design is unprecendented. It will be essential that the antennas at both the ground terminals and in the satellites be capable of generating nearly perfect orthogonal polarizations. The design problems to accomplish this differ considerably, however, depending on whether it's the antenna on the satellite or at the ground station. For the ground station antenna, it's essential to maintain the polarization purity only at the peak of the narrow pencil beam and over the tracking beamwidth. The real problem at the earth station is compensating for Faraday rotation resulting from the magnetic fields that encompass the earth. The orientation of polarized signals passing through the ionosphere is affected by these slowly varying magnetic fields which must be tracked and compensated for.

On the other hand, the spacecraft must maintain polarization purity over the entire shaped beam—about a 3 degree x 7 degree contour. To do this polarization, selective reflecting or transmitting surfaces consisting usually of either parallel grids of wire or grating strips are used. Because wideband feed designs are involved to provide the desired earth coverage, gain and coverage contours are not optimized for each individual frequency band."

AT&T, who probably has the greatest interest to develop additional bandwidth capacity, is also looking into using higher frequencies for satellite communications. But the question remains whether these bands are practical in view of the high signal attenuation that occurs in heavy rain. To improve upon the sketchy statistics available on rain depolarization and attenuation, the Comsat General domestic satellite for AT&T services will incorporate a centimeter wave beacon experiment. With this advanced all solid-state package, AT&T hopes to be able to determine more precisely the outages at an earth station from heavy rain; and what design margins are needed to overcome losses.

Maintaining isolation between overlapping channels

In selecting polarization diversity, both the RCA and Comsat/AT&T satellite systems need to provide an overall system polarization isolation of about 25 dB to maintain the required isolation between channels.

The budget for the AT&T system, Fig. 1 requires 35 dB polarization isolation at the earth stations and 33 dB at the satellite, resulting in 28 dB for each up and downlink and 25 dB overall. "This calls for a very sophisticated feed-design and a considerable hardware effect," explains Froome. "Independent rotation of transmitted and received polarizations requires a rather complicated waveguide system, while the polarization correction networks ($\Delta\phi$, ΔR) are based on recent theoretical work at BTL. As a result of the interference allocation, we're shooting for a 40 dB polarization isolation spec as a goal for the 30-meter antennas at

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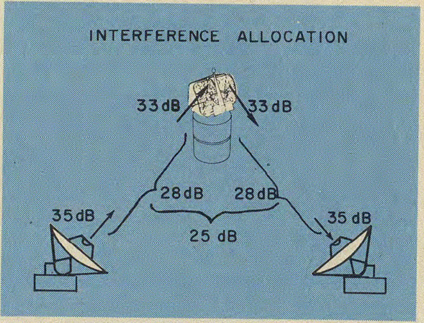
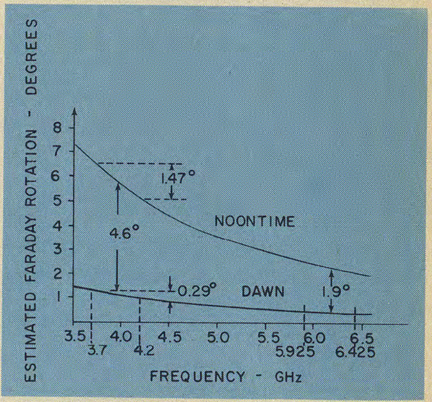


1. **To double capacity through frequency reuse,** an overall system polarization isolation of 25 dB is required. The interference from the three main sources are shown. Interference mechanisms in the antennas result from non-symmetries. In contrast, an Intelsat satellite only requires 14 dB of polarization isolation.



2. **Faraday rotation alters the orientation of linearly polarized signals** as they pass through the ionosphere. Maximum daily rotation is maximum during the day. It then subsides at cooler temperatures just before dawn. This graph does not include the effects of solar flares.

the ground terminal. We'll also have to be tracking the Faraday rotation." Since it is difficult to tell what rotation has taken place in the uplink, it was found necessary to monitor the misalignment in the downlink and intentionally transmit the signal to the satellite "crooked". The polarizations of the two signals must be received correctly because there is no way to compensate for the uplink Faraday rotation in the satellite.

Another problem in handling Faraday rotation arises from the fact that it is frequency sensitive, Fig. 2. "Our best theoretical estimates are that in the middle of the 4 GHz band, the orthogonal polarization can shift 4.6 degrees every day while at 6 GHz, the daily change is about 1.9 degrees," says Froome.

"If these effects were not compensated for, the results would be intolerable and would totally destroy the cross-polarization isolation." For example, interference due to angular misalignment is 20 log sin Θ. For a misalignment of only one degree, isolation drops to 36 dB and if the orthogonal signals are off by 4.6 degrees, only 22 dB of isolation is theoretically available. "It's, therefore, necessary that we track this misalignment to within 0.1 degree, by detecting and nulling the unwanted polarization."

To correct for polarization misalignment, Bell Labs has, therefore, developed a feed which incorporates a half-wave plate on a rotary joint controlled by a Faraday tracking system, Fig. 3. The Faraday tracking system is based on cross-correlation techniques. As the incoming vectors change, the half-wave plates shift to keep the signal aligned with the feed's output.

However, this is only part of the problem. "To obtain the 40 dB isolation, we need to tune the antenna," continues Froome. The vertical and horizontal signals as received are both eliptical and non-orthogonal. To clean them up, differential phase and attenuation elements ($\Delta\phi$ and ΔR) are used. Effectively they serve as conjugate

U.S. domestic satellites: what's up—what's going up?

Presently, there are two domestic satellite systems in orbit, Western Union's Westar and Telsat Canada's Anik. Each has two operational satellites in orbit. Both were built by Hughes Aircraft and are similar in design carrying 12 transponders with an average capacity of 7,000 two-way voice channels. RCA Globecom presently is leasing a transponder on Anik II and American Satellite Corporation is leasing two transponders on Westar I.

Both RCA Globecom and the AT&T/Comsat General domestic satellites, once they become operational, will provide double the channel capacity over their predecessors. These new entries in the U.S. domsat race are planned for launch in early 1976 and should begin service by late 1976. By using polarization diversity, 24 transponders are available, the equivalent of 14,400 voice channels.

The RCA Satcom System will consist of three 24 rf channel satellites in synchronous orbit and a series of earth stations to serve the U.S., Hawaii and Alaska. The RCA satellite Fig. A, being built by RCA's Astro Electronics Division, Hightstown, NJ, will use a three axis attitude-control system rather than the more common (and proven) spin stabilized system to keep the antenna pointing at the earth.

AT&T and GTE Satellite Corporation have petitioned the FCC to combine their satellite operations. Present plans include eight earth stations and three synchronous satellites providing coverage to the continental U.S., Alaska, Puerto Rico and Hawaii (Fig. B). The three satellites, being built by Hughes Aircraft will be placed in synchronous orbit by Comsat General, a subsidiary of Comsat, who then will lease the satellites to AT&T, and as proposed to General Telephone and Electronics Satellite Corporation (GSAT).

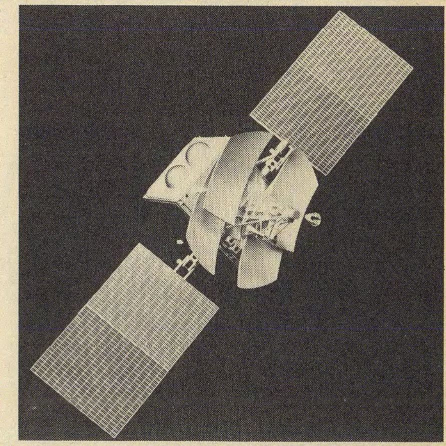
Th remaining entry in the U.S. domestic satellite plan is a proposed merger of Comsat General with IBM Corporation who wants to buy into the CML Saellite Corporation; now the property of Comsat, MCI Communications Corporation and Lockheed Aircraft. This merger is presently under scrunity by the Justice Department and the FCC. Western Union, among others, is contesting the venture on grounds that "the combination of Comsat General and IBM could forme an impenetrable barrier to competition."

Comsat AT&T satellite design

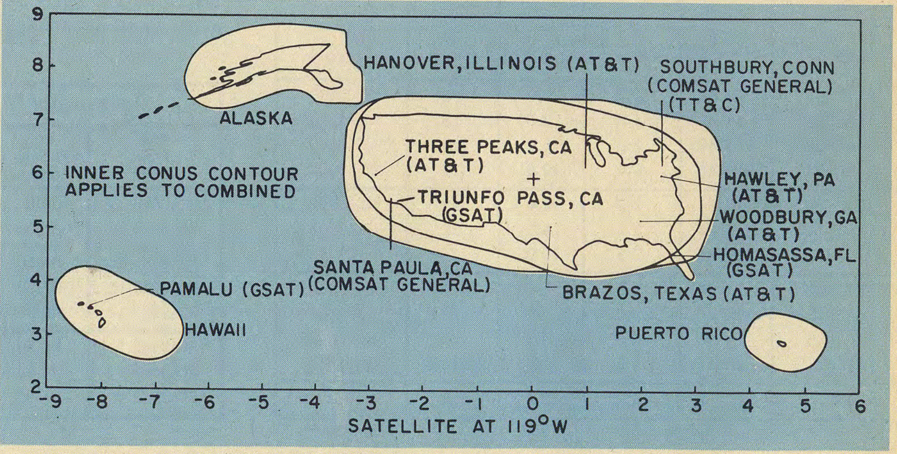
Like RCA's satellite, the AT&T/Comsat General domestic satellite consists of 24 transponders each operating in the 6/4 GHz band. When the system is fully loaded, 14,400 two-way telephone circuits will be available from each satellite. The diversity technique consists of transmitting vertical and horizontally polarized orthogonal signals to and from the satellite. All the vertical repeater channels receive from the combined Alaska and CONUS areas, while all the horizontal channels receive from the combined CONUS, Hawaii and Puerto Rico areas. Received signals are translated in frequency by 2225 MHz and retransmitted with the same polarization orienation. As shown in Fig. C, signals from the odd horizontal channels can be switched into either the CONUS or Hawaii spot beams. Similarly, the even horizontal channels can transmit to either CONUS or Puerto Rico. The even vertical channels can be switched in one of three ways: All power to CONUS, all power to Alaska or power divided between them. All power from the odd vertical channels is transmitted through the CONUS spot beam.

Two orthogonal (perpendicular) polarized offset parabolic reflector antennas are used measuring about 1.3 m x 1.8 m. Each has a multi-horn feed array to achieve the needed beam shapes for earth coverage and a polarization grating as shown on this months cover. The transmit gains (in dB) when added to the TWT power, yield a minimum EIRP of 33 dBW for spot beam coverage and 31 dBW for the combined service to Alaska and CONUS.

The vertically-polarized feed assembly consists of five offset feed horns and separate feed networks for transmit and receive. The feeds are mounted below to avoid blockage. Two feed horns are used to form the Alaska beam and three are used to form the CONUS beam. "The design of the vertically-polarized antenna was a bit more difficult than the horizontally-polarized antenna," explains Roger Rusch of Hughes Space and Communications



A. *RCA's three-axis stabilized satellite* measures approximately 5.3 ft x 4.1 ft x 4.1 ft and its orbiting weight is about 1,000 lbs. A telemetry horn antenna appears at right.



B. *AT&T's domestic satellite system* will consist of three stationary satellites (two operating and one for backup or for sun transit). The five AT&T earth stations are located at Hawley, PA; Woodbury, GA; Hanover, IL; De Luz, CA and the San Francisco area. GTE Satellite Corporation (GSAT), who will use some of the communications capacity, will have terminals in Hawaii, Florida and near Los Angeles. The telemetry tracking and command (TT&C) ground stations on the East and West Coasts are run by Comsat General.

matches to the imperfections inherent in either the satellite or earth station antennas.

The feed system, Fig. 3, being built by Ranatec Corporation, Calabassas, CA, does introduce a slightly increased insertion loss of about 0.2 dB—over systems without polarization diversity. For example, an Intelsat ground terminal feed without polarization equipment has about a 0.2 dB insertion loss operating into a 17-degree K low noise receiving system. This results in an overall station temperature of approximately 60°K. "The additional 0.2 dB will increase the overall station temperature to 79 degrees K (at a 20 degree elevation angle)," says Froome. This results in a slight degradation in G/T, "but we elected to take the loss in exchange for the cross-pole. We think the price is right when you consider we've doubled channel capacity."

Bell Labs has tested the feed at Hughes Aircraft's cassegrain antenna facility in Glenwood, AR, and it has obtained some fairly encouraging results. The on-axis cross polarization is better than 40 dB, well within specifications. The cross-pole lobe structure allows a tracking error of more than ±0.016 degrees for the earth station's narrow pencil beam (3 dB beamwidth is 0.16 degrees at 4 GHz). This is important in case the earth station's 30 m antenna is moved off axis; for example by strong winds.

Interleaved channels important

Another factor on why it's possible to achieve this polarization isolation between signals is the manner in which the channels are interleaved. According to Roger Rush of Hughes' System Engineering Department, El Segundo, CA, the spectrum is separated into 24 channels, 12 vertically polarized and 12 horizontally polarized, Fig. 4. Each 36 MHz channel is spaced 40 MHz apart resulting in 6 MHz guard bands.

"The guard band compliments the polarization isolation inherent in the antenna design," says Rusch. "For example, an overall noise budg-

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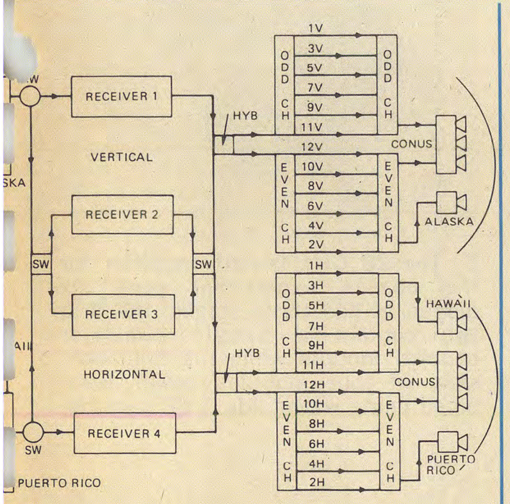
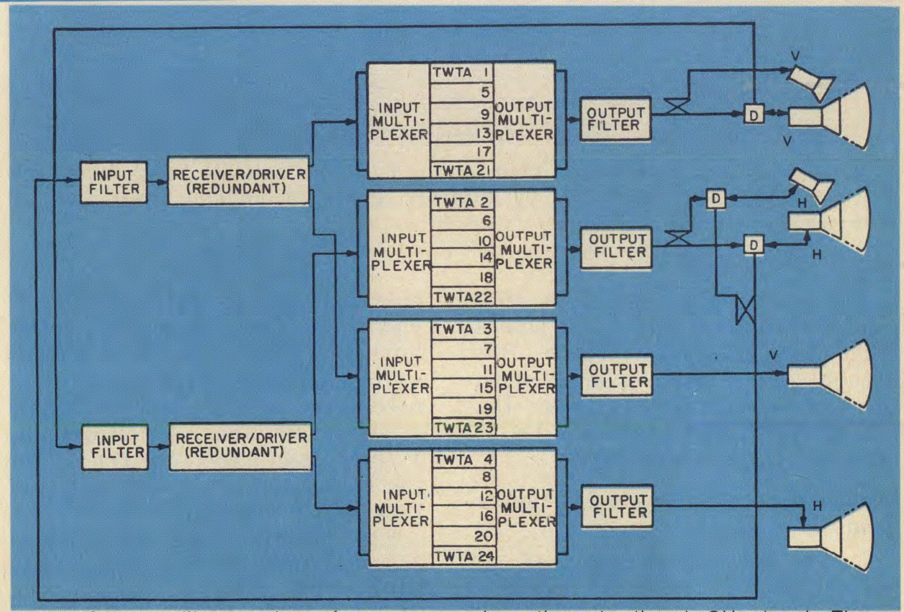


C. *In the Hughes' satellite for Comsat/AT&T*, any two of four TDA receivers may be used (receivers 1, 2 or 3 with horizontal channels and receivers 4, 3 or 2 with vertical channels). Twelve 5 W TWTAs are used for the horizontal repeater and twelve 5.5 W TWTs for the vertical repeater (because of the lower gain from the vertical antenna).



D. *RCA's satellite consists of two receive ports and four transmit ports.* The receive ports, one for each polarization, amplify the two groups of odd and even-numbered channels and heterodyne them to the 4 GHz band. The channels are then separated, amplified in TWTs, and then collected into four groups of six channels by output multiplexers.

munications Group, El Segundo, CA. Essentially its a matter of not being able to place them in an optimum position in front of the reflectors because of lack of space. For example, the CONUS beam requires three feed horns but unfortunately, the optimal positions for the feeds overlap due to their physical size. "As a result, we get suboptimal performance on the vertical antenna. Because the CONUS horn array illuminates the reflector with less efficiency, transmit gain for the vertically-polarized antenna is 26.5 dB at 4 GHz vs 27 dB for the horizontally-polarized antenna. To handle this problem, Hughes' engineers have used 5.5 W TWTAs in the vertical channels vs. 5 W tubes in the horizontal channels. For receive, the vertical antenna has a gain of 24.5 dB because it has more area to cover.

The horizontally-polarized feed assembly consists of six feed horns and separate feed networks for transmit and receive. These feed horns are arranged to form beams for CONUS, Hawaii and Puerto Rico.

The polarization grids, which consist of conductive strips on a dielectric in front of the reflectors, can, in practice, reduce the polarization interference level to −33 dB. The strips are one inch wide and 0.5 inch apart. They are mounted to a rigid frame which is bent to form a 50 inches by 70 inches inside diameter.

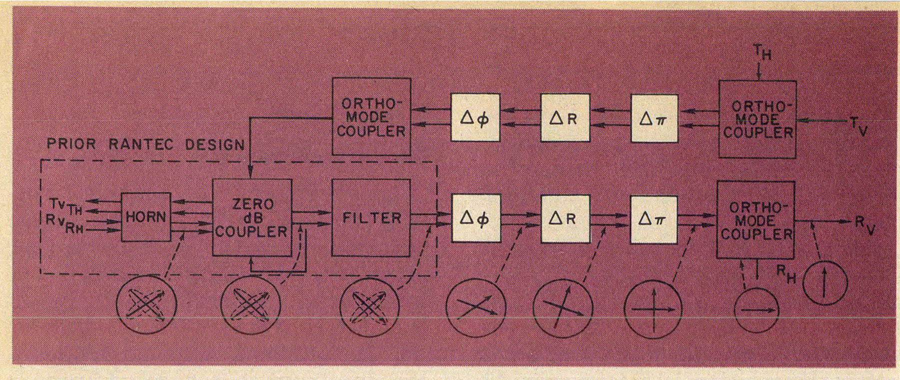
RCA's design

The antenna compliment on the RCA satellite, Fig. D, consists of four separate grated reflectors with offset horns mounted on the earth-facing platform. Two of the antennas generate vertically-polarized beams covering Alaska and the continental U.S., the other two produce identical horizontally-polarized beams. According to Robert Miller of RCA's Commercial Sat Com Group/Astro-Electronics Div., Hightstown, NJ, one of each pair operates at 4 GHz only while the other uses a duplexer for transmit operation at 4 GHz and receive at 6 GHz. In addition, feed horns placed off axis on the two western antennas produce beams toward Hawaii.

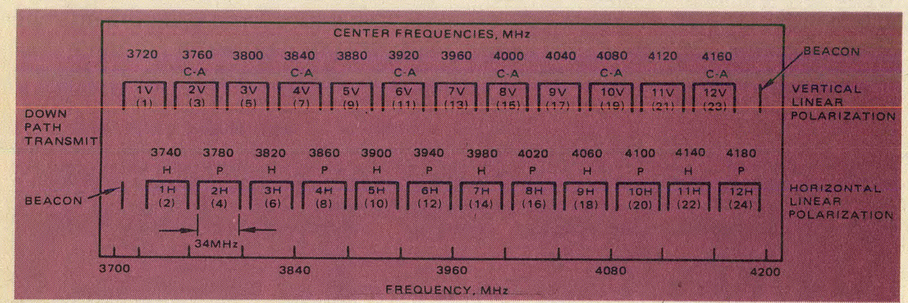
Unlike the Hughes design in which polarization grids are in front of the reflectors, the RCA reflectors are paraboloids of revolution whose surfaces comprise parallel wires embedded in fiberglass. The reflected waves are thus polarized along the wire direction. In addition to the gratings being virtually transparent to the orthogonally-polarized wave, the displacement of the feeds are both lateral and axial, providing the high polarization isolation required. The common parallel axes of all four reflectors are elevated approximately six degrees from the equatorial and the major axes of the reflectors are rotated by 20 degrees in a clockwise direction (looking toward the earth).

Lightweight graphite fiber epoxy components are used for the RCA transponders. "This gives a tremendous weight advantage," claims Miller. "An invar filter weighs about 2.1 lbs vs 0.5 lbs for one using graphite yet their performance is comparable. Also, because a three-axis attitude control is used, the solar arrays rather than being cylindrical consist of two rectangular panels each 7.15 sq. ft continously oriented toward the sun. "This means that only one-third the usual number of the silicon solar cells are required to power the 24 transponders," says Miller. As a result, a lower cost launch vehicle is planned.

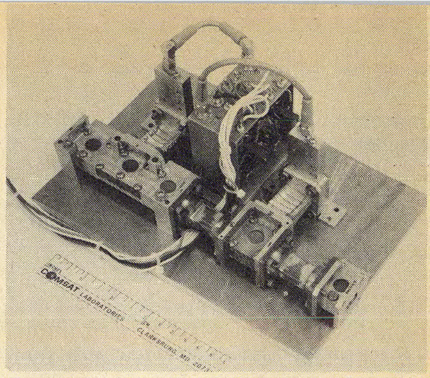
In the contiguous U.S., RCA has located earth stations near New York, San Francisco and Los Angeles. Additional stations are planned near Dallas, Atlanta, Seattle, Denver, Chicago, Washington, DC and Hawaii. Because of Alaska's topographic features, 16 earth stations are planned there. RCA is also preparing to construct and operate receive-only stations primarily for television or other broadband services. □



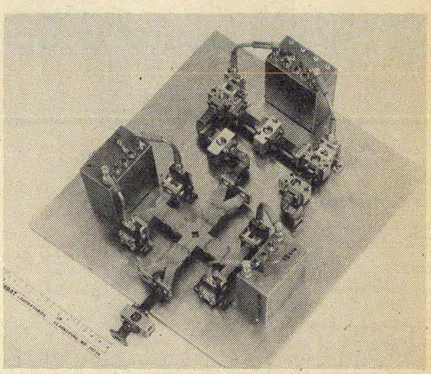
3. **The dual-polarized feed system of the AT&T ground station** uses a $\lambda/2$ plate to align the incoming vectors (misaligned from Faraday rotation). To correct antenna assymetrics, differential $\Delta\phi$ and ΔR elements are used to clean up the ellipticity and non-orthogonality of the signals received, resulting in an isolation of 40 dB.



4. **Frequency and polarization plan** for the 6/4 GHz satellite band shows how frequency spectrum is reused by interleaving and orthogonal linear polarization.



5. **Two-stage 19 GHz amplifier** uses Impatt diodes imbedded in circuit modules and coupled to reduced height waveguide to optimize dc-rf efficiency. Each module is coupled to its circulator via a seven section stepped impedance transformer.



6. **The 28 GHz Impatt amplifier** for the beacon experiment uses six diodes. It has three stages; the first two contain a single circulator coupled module while the final consists of four modules power combined using waveguide 3 dB hybrids.

et for the entire systems is 4800 pW. Fortunately, only 1/2% comes from cross-pole interference. But without the 6 MHz guard bands, the cross-pole contribution to the overall noise budget would be at least several hundred pW."

Beacon experiment evaluates higher bands

To test out the feasibility of using the higher bands for satellite service Comsat Labs, Clarksburg, MD, has designed and built two beacons that transmit at 19 and 28.6 GHz.

Their carrier frequencies are derived from a stable-master oscillator at 132.2 MHz and then multiplied up. Specifically, the signal is amplified, frequency doubled and fed to two cascaded transistor triplers, which multiply the frequency to 2.38 GHz. A transistor amplifier boosts the level to +27 dBm (two such oscillator chains are included for redundancy). A hybrid permits either oscillator chain to drive both the 19 and 28 GHz transmitter chain, but only one oscillator operates at a time. In the 19 GHz chain, a two-stage Impatt amplifier (Fig. 5) is used amplify the signal to +29 dBm. In the 28 GHz chain, a three-stage Impatt amplifier (Fig. 6) is used the level from 8.5 to 29.5 dBm.

The two solid-state transmitters are designed to operate for at least two years using excess power available during the early spacecraft operation. The beacon that transmits the 19.04 GHz uses a diode switch to transfer the signal alternately to vertical and horizontal polarized antennas. An ultra-stable 1 kHz oscillator controls the the switching speed. The EIRP of this beacon is 21 dBW.

The other beacon transmits at 28.56 GHz also with 21 dBW EIRP. This carrier is phase modulated to produce two first-order sidebands, 7 dB down at ±264.4 MHz (for the first two satellites) and ±528 MHz (for the third satellite once it is launched).

Two sets of linearly-polarized antennas are used on the satellite. One set is linearly polaried at 28 GHz. The other antenna set is capable of dual orthogonal polarization at 19 GHz. This dual polarized single aperture antenna is essential for phase-coherence measurements planned by Bell Labs. Each set is an array of two linearly-polarized pyramdial horns that are tilted to rotate the polarization angle. A magic tee divides the input power equally to the horns and a grating in the aperture of each horn insures polariaztion purity.

At the ground station located at Bell Lab's Crawford Hill Facility in NJ, the received power will be −98 dBm. The stable master oscillator allows these faint signals to be measured by the receiver because the bandwidth is only 100 Hz at 19 GHz and 150 Hz at 28.6 GHz. Short-term stability is such that 90% of the carrier power is contained in these bands. Long-term drift is less than 1 ppm.

According to Don Cox of Bell Labs Satellite Systems Research Department, Holmdel, NJ, the seven-meter dish being used to recover these signals has a surface tolerance that makes it suitable for frequencies over 100 GHz. This feature allows the antenna to be used for radio telescope purpose as well.

When the beacons are in operation, data will be gathered to determine earth station outage on a statistical basis. Heavy rain, 100 mm/hour or more has been found to introduce 50 dB total attenuation at millimeter frequency, more than enough to knock out any ground terminal. Besides attenuation, depolarization, bandwidth coherency and diversity, effects will also be studied. AT&T is encouraging institutions, universities, etc., to participate in this experiment to provide a good statistical base over diverse climatic regions. ••

r&d roundup

Harvey J. Hindin, Contributing Editor

Smith does it again

One would think that new uses for the Smith Chart, which has been around since 1939, would have been exhausted by now. It's been the subject of countless articles including a comprehensive book by Smith himself. Nonetheless, there are still things that the chart can do—and they are not obscure, trivial or of little interest particularly for transmission line analysis.

The latest use of the Smith Chart developed by R. M. Arnold of Bell Labs is discussed in "Transmission Line Impedance Matching Using The Smith Chart" (*IEEE Trans. on MTT*, Vol. 22, No. 11, pp. 977-978, November, 1974).

Arnold's detailed analysis and illustrative example finds the characteristic impedance and electrical length of a transforming section which converts a complex load impedance to match a real source impedance. Since we do not know the matching characteristic impedance, we cannot plot the load impedance on the chart by the usual methods. The graphical search procedure is a clever way around the problem.

A barrel full of microwaves

Permittivity measurements on dielectric rods at microwave frequencies when done in the classical manner, suffer from having to accurately machine the sample to fit snugly in the closed resonator. Otherwise, air gaps distort the measured resonant frequency. One solution to this problem is to use an open structure, Fig. 1. The spherical-mirror device (a) is ideal for flat sheet samples while for rod shaped specimens the barrel configuration (b) is appropriate.

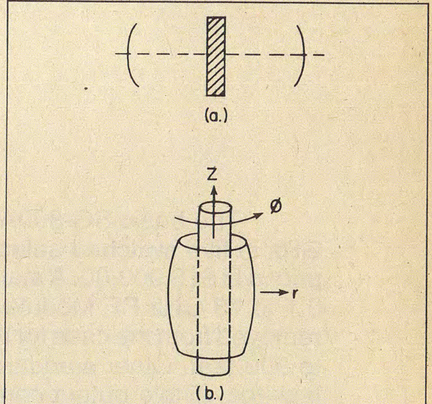
A complete analysis of the resonant frequencies and the cylindrical sample dielectric permittivity has been performed by P. Nagnethiram and A. L. Cullen of University College in London and presented in "A Microwave Barrel Resonator for Permittivity Measurements on Dielectric Rods" (*Proc. IEEE*, Vol. 62, No. 11, pp. 1613-1614, November, 1974).

The authors derive expressions for the desired resonant frequencies for the empty barrel in both E and H modes by two separate methods. In the first technique, the Hermite function form of the axial field variation is obtained from circular waveguide modes using a Fourier integral. The second method uses an approximate solution to the wave equation in prolate spheroidal coordinates, which is then used in a variation method to derive a resonance formula. Unfortunately, neither method is detailed.

The estimated accuracy of both methods is about 0.15%. Twenty-four E modes and 19 H modes are experimentally examined in X-band to verify the predictions. Comparison between theory and experiment is excellent. Equations for the dielectric-loaded barrel are then derived by matching field components across the air dielectric interface. Again, only the results are given.

These transcendental equations are then used to predict rod permittivity from known physical dimensions of the barrel system and the measured resonant frequency with the dielectric in place. Typically, for a dielectric with permittivity of 2.52 at 10 GHz, measured values for different diameter rods differed by about 0.01. The E_{050} mode is used.

Nagenthiram and Cullen close their paper with an error analysis and a discussion of the experimental techniques used to take their data. They also note that the same method may be used to make loss tangent measurements.



1. The barrel-shaped open resonator (b) is more convenient for examining cylindrical samples than the spherical mirror structure (a). Both structures have high Q values such that modes off resonance are highly attenuated.

The above mentioned references are available from IEEE, Single Sales Publication, 345 East 47 Street, New York, NY 10017.

MICROWAVES

HIGHLIGHTS OF THIS SECTION:

What's behind the design of a portable spectrum analyzer?...p.36

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TEST & INSTRUMENTATION SECTION

Low cost counter offers 1 Hz resolution...p.43

Behind The Design Of A Portable Analyzer

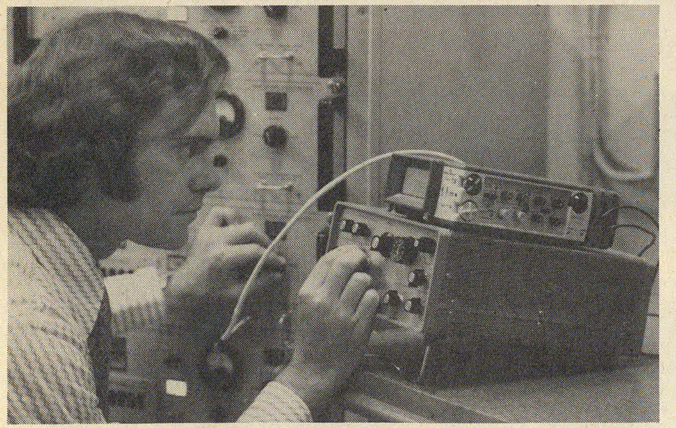
Here is a review of some of the important design problems encountered in building a 2 to 25 GHz Yig-tuned spectrum analyzer and how they were handled.

IMPROVEMENTS in Yig-tuned filters and MIC rf devices have made possible compact wideband spectrum analyzers that rival the performance of units which are considerably greater in size and weight. A case in point is the spectrum analyzer shown in Fig. 1. The unit measures only 5x9x13 inches and weighs 22 pounds, yet provides features such as preselected frequency coverage from 2-25 GHz, 70 dB dynamic ranges, video filtering or pulse stretching for signal enhancement, in addition to the customary sweep and tuning controls normally found on much larger analyzers.

The analyzer was developed for applications where broad frequency coverage, portability and preselection were of prime importance, such as for spectrum surveillance, emi/rfi measurements, and field installation and testing of microwave links or radar transmitters. By choosing the receiver tuning range from 2 to 25 GHz, the majority of radar, satellite communication and terrestrial microwave frequencies in use today or planned future are covered.

This article is not intended to be a tutorial on spectrum analyzer design. It does provide some insight into one solution to several key problems associated with the design of a portable, broad frequency coverage, preselected microwave analyzer/receiver.

Bob Hoss, Project Manager and **Frank Reisch,** Manager, Applied Research Department, Telecommunication Systems Division, Collins Radio Group, Rockwell International, Dallas, Texas 75207.



1. **Compact spectrum analyzer** provides preselection over a 2 to 25 GHz band yet measures only 5 in. x 9 in. x 13 in. and weighs 22 lbs.

Features of the receiver

Table 1 summarizes the characteristics of the analyzer. The unit can be either manually tuned or swept over the 2-25 GHz frequency range in 1 GHz increments. The reasoning behind this choice of tuning increments and other features will be apparent later. As shown in Fig. 2, the preselector output is mixed with either the fundamental (in the range from 2-8 GHz) or third harmonic (in the range from 8-25 GHz) of the local oscillator chain output, to produce the first i-f of 425 MHz. A second conversion to 21.4 MHz permits the use of commonly available crystal filters to achieve bandwidths as narrow as 5 kHz. (However, a 50 kHz bandwidth provides adequate sensitivity for the intended applications).

Both am and fm detectors are provided for carrier demodulation. This feature can be useful for many purposes, as for example, to positively identify a signal source when conducting an rfi survey. An operational mode is also included which permits phase locking the receiver local oscillator to an internally-generated reference signal to improve short term stability and reduce the incidental fm inherent in yig-tuned oscillators. Achieving good performance up to 25 GHz requires special attention to problems in three areas—Yig preselector design, the mixing scheme, and tracking of the preselector with the local oscillator.

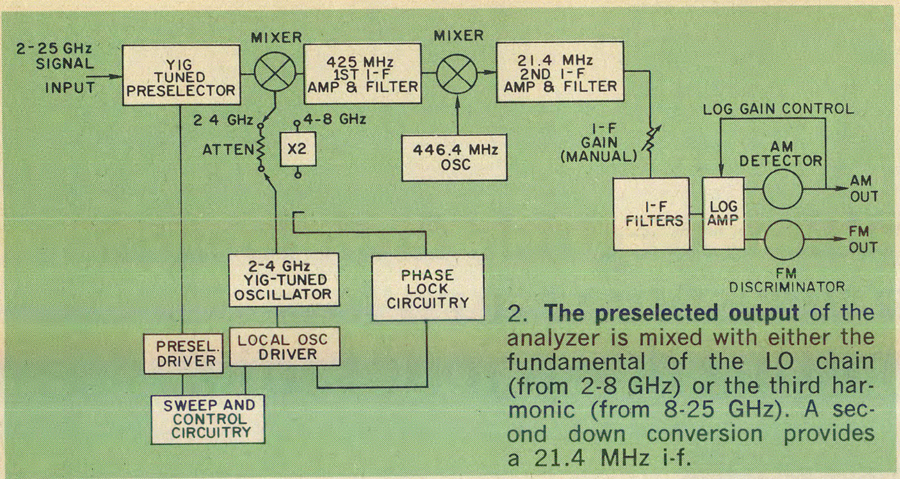
Preselector design

A Yig-tuned preselector permits a spurious free

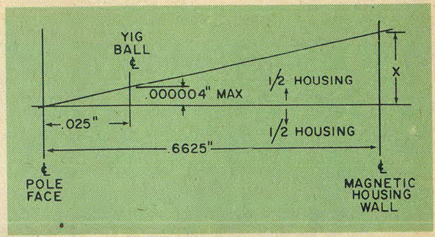
(continued on p. 38)

Table 1. Performance Characteristics

Tuning range	2-25 GHz in 1 GHz Bands
Resolution	50 KHz, 500 KHz, 5 MHz
Log IF range	≃ 60 dB
Dynamic range	70 dB
Image and spurious rejection	≃ 60 dB
Video processing	UNFILTERED 2.5 MHz FILTERED 10 KHz STRETCHED 100 μSEC
Dispersion	FIXED 1 GHz OR VARIABLE 0-250 MHz
Sensitivity (Typical)	−100 DBM 2 GHz TO 8 GHz −95 DBM 8 GHz to 25 GHz
Power required	115 OR 220V (+10-20%) 47-420 Hz 60W NOM
Size	5" x 9" x 13"
Weight	22 LBS



2. **The preselected output** of the analyzer is mixed with either the fundamental of the LO chain (from 2-8 GHz) or the third harmonic (from 8-25 GHz). A second down conversion provides a 21.4 MHz i-f.



3. **The Yig-filter poles** can be ground flat to within $\pm 0.4 \times 10^{-5}$ inch by establishing a grinding plane to within 0.001 inch as shown.

display that simplifies signal identification. Using a Yig-preselector, however, does present some difficulties particularly in good linearity, low hysteresis and the tracking of the Yig resonators over the full 2-25 GHz frequency range.

For uniaxial ferrite Yig spheres oriented with their > 111 > axis aligned with the externally-applied magnetic field, the frequency of resonance is given by:

$$f_{res} = \gamma \, (H_o + H_A) \qquad (1)$$

where γ = The material gyromagnetic ratio
H_o = The extremely applied field strength
H_A = The crystal internal anisotrophy field.

Yig material has a negligible anisotropy field ($\simeq 45$ gauss), therefore to tune to 25 GHz requires approximately 9000 gauss external field. The first step is then to achieve a magnet design that permits the attainment of that field strength with good linearity and low hysteresis.

For this analyzer design, the best magnet housing material that could be found was an alloy consisting of 47% nickel and 53% iron. This material has a saturation flux level (B_{max}) of 15,000 gauss and a coercive force (H_c) of less than 0.03 oersteds. Initially, there was some concern that adequate linearity could be achieved with a pole gap width that would permit a reasonable size rf structure, since large gaps result in undesired saturation effects with a subsequent loss of tuning linearity. After some experimentation, it was concluded that a linearity of better than $\pm 0.1\%$ overall (2-25 GHz) could be achieved with a pole gap of 0.030 in.—a spacing just sufficient for the rf structure.

If a larger pole gap had been necessary, a magnet housing material having a higher B_{max} would be required. Unfortunately, the materials having a higher B_{max} also have greater hysteresis. For example, a compound consisting of 2% vanadium, 49% cobalt and 49% iron offers a B_{max} of 23,000 gauss, but has an H_c which varies from 0.18 to over 1 oersted depending on how it is annealed. This is unacceptable for filter applications.

While even $\pm 0.1\%$ is still rather large, by tuning the receiver in 1 GHz steps, the tracking errors due to nonlinearities and hysteresis are greatly reduced, and were found to be ± 3 MHz for linearity and 8 MHz maximum for hysteresis, over any 1 GHz segment.

The second area of concern deals with resonator mistracking. The ability of the two resonators to track each other over the full tuning range of the filter is related to the parallelism of the pole faces (i.e., the uniformness of the magnetic field). Calculations show that the pole faces must be parallel to within 0.000008 inches in order to keep mistuning to less than 10 MHz over the 2-25 GHz frequency range! This may be determined as follows: For a magnet with a large air gap:

$$H = \frac{0.4\pi NI}{l_g} \qquad (2)$$

where H = the magnetic field strength (gauss)
l_g = pole gap (cm)
N = number of turns on magnet coil
I = tuning current (amps)

Also, the frequency at which the yig sphere resonates is approximately

$$f_o = 2.8H \qquad (3)$$

where f_o = the resonant frequency in MHz

Substituting Eq. (2) into (3) yields

$$f_o = \frac{3.519NI}{l_g{}^2} \text{ (MHz)} \qquad (4)$$

To determine the sensitivity to change in pole gap, the derivative of Eq. (4) is obtained with respect to l_g;

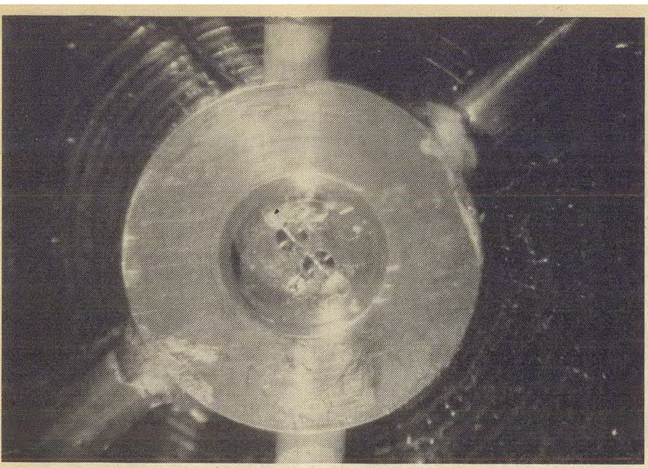
$$\frac{df_o}{dl_g} = \frac{-3.519NI}{l_g{}^2} \qquad (5)$$

Measuring a variation of 0.000008 in. is difficult unless interferrometer techniques are used. To avoid this, an indirect method is possible. By turning the filter housing about a centerline through the poles, and indexing off the walls of the housing, a grinding plane can be established to better than 0.001 in., which is the accuracy required to assure that each pole face is that to within 0.000004 in. As shown in Fig 3, this can be accomplished by using similar triangles:

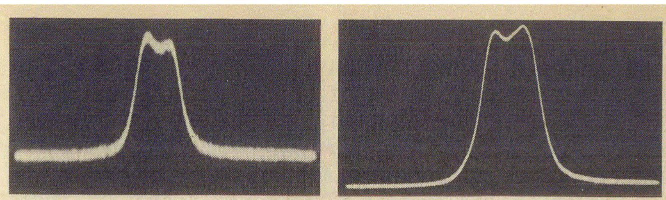
$$\frac{X}{0.6625} = \frac{0.000004}{0.025} \text{ ; } X = 0.00106 \text{ in.} \qquad (6)$$

This 0.001 dimension is easily measured.

The rf portion of the filter is shown in Fig 4. Based on an overall receiver maximum bandwidth requirement of 5 MHz, and considering all sources of preselector /LO mistracking, a minimum preselector bandwidth of 25 MHz was found necessary. The design procedure of Ref 1 was followed to establish the necessary coupling loop sizes for a 30 MHz minimum design bandwidth. Insertion loss of the final filter typically measured 3-4 dB and its bandwidth varied from 30 MHz at 2 GHz to 40 MHz at 25 GHz. Figure 5 shows typical filter passband shapes at 2 and 26 GHz. The key to obtaining satisfactory rf performance to 25 GHz is in keeping



4. **RF portion of the Yig filter** shows minute Yig spheres and the rf input band output ports.



(a) (b)

5. (a) **Noisy filter passband** at 2 GHz results from lower signal generator level used to avoid limiting in filter. The 3 dB bandwidth is 30 MHz. (b) The 3 dB passband at 25 GHz is 40 MHz.

circuit physical size to an absolute minimum.

A trade-off among size, weight, and power consumption was made with the result that the final filter design tunes to 25 GHz with less than 500 mA, dissipates less than 1.5 W at 25 GHz, weighs 1 lb, and is approximately 2.4 in. long by 1.5 in. diameter.

Mixer /LO design

Since the initial design objective for the receiver was small size and light weight, the trade-off between using fundamental mixing over the entire frequency range, and harmonic mixing over the higher bands, weighed heavily in favor of harmonic mixing. To use fundamental mixing only would have required a multiplicity of Yig oscillators and switches, which was avoided with the harmonic approach.

The next step was to obtain a suitable broadband mixer. Neuf[3] describes a double-balanced mixer, having broadband balun transformers on both the RF and LO ports, which permits the achievement of low conversion loss ($\simeq$8dB) over the full 2-25 GHz frequency range. Since a balanced mixer cannot be used as an even harmonic mixer, a times two frequency multiplier, Fig. 6, is needed in the LO chain to provide a 4 to 8 GHz output from the 2 to 4 GHz Yig oscillator. The resulting mixer /LO configuration shown in Fig 2 provides fundamental mixer operation from 2-8 GHz and third harmonic operation from 8-25 GHz.

The final receiver front end design yields a tangential sensitivity (using 50 KHz IF bandwidth) of −100dBm, typically, from 2 to 8 GHz, and −95dBm from 8 to 25 GHz. The maximum signal level into the receiver before saturation is −30dBm, yielding a 70dB dynamic range. A 70dB range i-f attenuator is provided, as well as an i-f long range of 60 dB.

Preselector and Local Oscillator Tracking

Besides the design of the 2-25 GHz preselector and mixer/LO chain, the third major consideration in this design is to minimize the mistracking between the preselector and local oscillator. If appreciable mistracking occurs, a loss of sensitivity results.

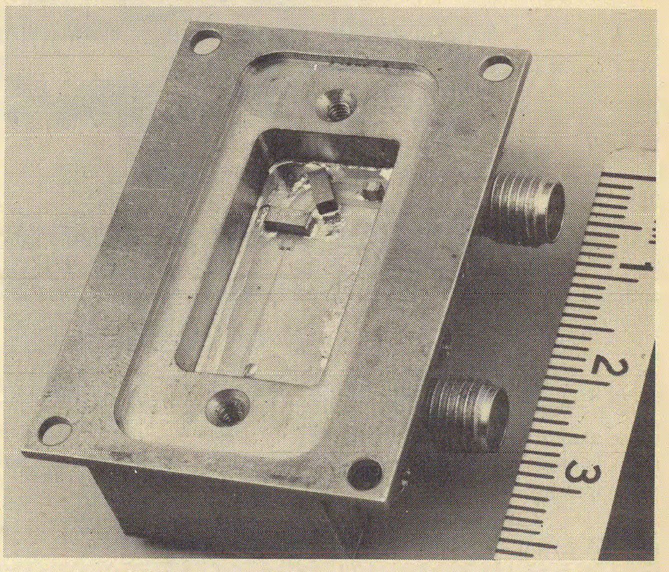
Magnetically tuned devices inherently have, depending on material and tuning range, some nonlinearity, hysteresis and retentivity, and a relatively slow response time due to magnet coils inductance. There is also a problem of drift with temperature due to both self heating and changes in ambient temperature. It is desired that this analyzer operate over a range of −20°C to +55°C, creating a need for stiff controls on frequency drift with temperature. The drift mechanisms are numerous. The Yig devices drift with temperature due to variations in magnet coil resistance, pole gap, and the Yig material itself. Voltage references, component values and semiconductor devices in the driver circuitry also vary with temperature and must be carefully considered in the design of the driving circuitry.

Aside from minimizing preselector and local oscillator linearity, hysteresis problems by proper magnet design and temperature drift by employing heaters if the temperature range is extreme, there are a number of circuit design approaches that may be taken to alleviate the tracking problem over frequency and temperature. In summary, they are:

- Limit the maximum frequency tuning range per band.
- Use low or zero temperature coefficient voltage references.
- Use of large amounts of current or voltage feedback.
- Sense ambient temperature changes and provide compensation.
- Provide dynamic hysteresis compensation.
- Limit power dissipation and provide ample heat sinking.

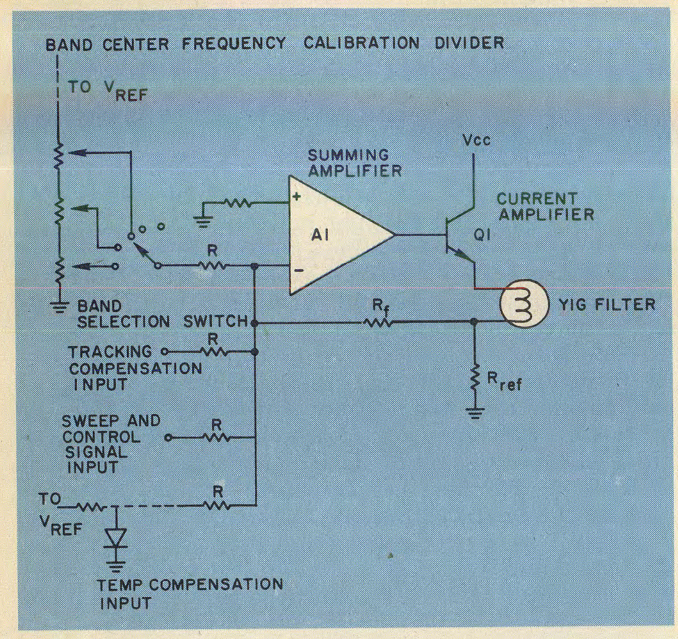
By tuning the receiver in 1 GHz steps, the linearity and hysteresis errors are considerably reduced. A precision, low temperature coefficient resistive divider chain establishes the center frequency of each band, thus piecewise approximating the YIG devices tuning curves.

Currents through the Yig devices are sensed by precision low temperature coefficient resistors in series with each device, and feedback to the driver



6. **Slotline/coplanar doubler** has a conversion loss of only 3.5 dB to 4.9 dB over the 2-4 GHz band.

op-amp minimizes drift due to magnet coil resistance changes. All active devices have been chosen for their low drift characteristics and large amounts of feedback have been used to stabilize the operating point. All critical summing and reference resistors and voltage references have a temperature coefficient of 1 ppm or less. All trimpots are low temperature coefficient cermets.



7. **Yig-driver circuitry is designed to minimize drift.** The summing and reference resistors and voltage references have a temperature coefficient of 1ppm or less.

Ambient temperature compensation is provided by a temperature controlled voltage source which uses a silicon diode as the temperature sensing element. A diode offers a relatively linear change in forward voltage with temperature. The sensor is mounted in the rf module with the Yig preselector.

Power dissipation has been limited to 3-5 watts per module. All major power dissipation devices, such as the Yig device driver transistors and voltage regulators, are attached to the receiver main frame whenever possible, rather than being included within a module. To compensate for dynamic hysteresis, a dual section potentiometer is used for the sweep rate control, with the second section supplying a correction voltage to the preselector which varies as a function of sweep rate.

Figure 7 is a simplified schematic of the YIG driver circuitry which incorporates these features. By the application of all these techniques, the mistracking over a $-20°$ to $+55°C$ temperature range is reduced to the order of ± 10 MHz. ••

References

1) W. Venator, "Charting a Simpler Course to the Design of YIG Filters," Electronics, pp. 118-126, (Mar. 3, 1969).
2) J. K. Hunton and J. S. Takeuchi, "Recent Developments in Microwave Slot Line Mixers and Frequency Multipliers," 1970 Int'l Microwave Symposium Digest, pp. 196-199.
3) D. Neuf, D. Brown, and B. Jaracz, "Multi-Octave Double-Balanced Mixer," Microwave Journal, (January, 1973).

Test your retention

1. Why must mistracking between preselector and local oscillator be kept to a minimum?
2. What are the sources of tracking error?
3. Why is a doubler used in the LO chain?

APPLICATION NOTES

Hall effect helps magnetic measurements

Gaussmeters provide a convenient way to measure the flux density of magnetron assemblies, TWTs or other devices where a permanent or alternating magnetic field is encountered. The heart of these complex instruments is a tiny semiconductor chip acting as a Hall-effect generator.

The basic premise of a Hall generator is simple. When a magnetic field is applied to a semiconductor that is carrying a current, a voltage is developed at right angles to the current flow. Several semiconductor materials are suitable for Hall generation, but Indium Arsenide seems to have outstanding properties, including an extremely low temperature coefficient (approximately 0.04% per degree centigrade).

When using a gaussmeter, it's vital to select the proper probe. Hall-effect generators are packaged in several basic probe styles: flat, transverse, tangential and axial, for example. Two independent probes may be combined to form a differential probe, which is used for simultaneously measuring gradients in a field at two different points.

"Wide Range Guassmeters," a seven-page publication from RFL Industries, Inc., provides some interesting tips about selecting and using these little-discussed instruments. Although heavily laden with products, the brochure gives some insight to the principles and practice of flux-density measurements. **RFL Industries, Inc., Instrumentation Division, Boonton, NJ 07005.**

CIRCLE NO. 204

Are your standards up to snuff?

How often should your laboratory standards be calibrated? Knowing the required measurement accuracy and oscillator drift rate or stability, you could probably find a high-order equation to determine the proper calibration period. Or, you can simply use a nomograph. A handy three-column design-aid is available free, from: **True Time Instrument Co., 429 Olive Street, Santa Rosa, CA 95401 (707) 528-1230.**

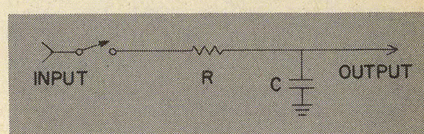
CIRCLE NO. 205

APPLICATION NOTES

Signal buried in noise? Try a boxcar integrator!

It is often necessary to reconstruct an entire waveshape of a repetitive signal buried in noise, or to measure the average value of a small segment of a repetitive signal. There are many sophisticated ways to make these measurements, but one of the most efficient and economical means is by a method called "boxcar signal integration."

A boxcar integrator is essentially a signal processor which uses a narrow filter to reduce the noise with little or no effect on the signal bandwidth. A simple integrator (see Fig.) consists of a gated switch and a low-pass filter. During the time when the gate is closed, the repetitive input signal is applied to the low-pass filter which acts as an integrator. After several pulses have been gated into the low-pass filter, the circuit's output is the average of the signal plus noise when the gate was closed. Since most noise has an average value of zero, the signal-to-noise ratio is improved.



The boxcar integrator is essentially a switched low-pass filter.

In its most fundamental practical form, the boxcar integrator is made up of a sampling oscilloscope, a few capacitors and a voltage-recording instrument. The sampling scope can be used with or without a display, and the voltage recorder can be either an X-Y recorder, an oscilloscope with camera or a storage oscilloscope. The basic principle is the process of controlled sampling and filtering of a repetitive waveform. By setting the rate at which the waveform is scanned and selecting the filter constants, the amount of signal integration can be carefully controlled.

The concept of boxcar integration can be used on almost any repetive signal to reduce or eliminate noise. An application note on the subject, No. 44L1.0 is available from: **Tektronix, Inc., P. O. Box 500, Beaverton, OR 97005 (503) 644-0161.**

CIRCLE NO. 203

Cost-conscious counter provides 1 Hz resolution to 4.5 GHz

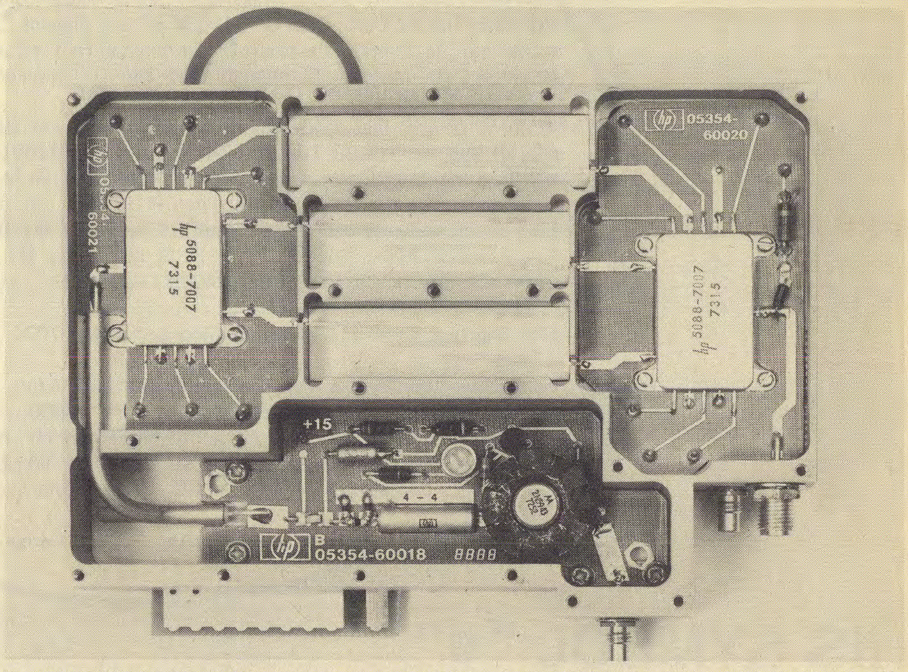
Hewlett-Packard's new 5341A automatic frequency counter offers several original ideas designed to provide a high level of performance at a modest cost. For example:

- Thin-film switchable filters are used to select comb lines for heterodyne mixing. Since the cost of this new technique is proportional to the number of filter sections, the 4.5 GHz counter sells for about \$3,600, while a 1.5 GHz model (option 003) sells for about \$1,000 less.
- Diagnostic aids allow the user to pinpoint the sources of internal malfunctions directly from the front panel.
- A choice of automatic or manual modes provides an option of searching the entire frequency range for the lowest frequency signal or restricting the search to one of ten relatively narrow frequency bands.
- Compatibility with ASCII codes means that no ancillary equipment is required to tie in with the HP Interface Bus System.

The 5341A's design is a departure from the rest of HP's automatic counter line since it is based on a heterodyne principle instead of a transfer oscillator technique. Mid-band tolerances specifications of ±250 MHz (max deviation; 0 to 500 MHz, 1.0 to 4.5 GHz) and +125 MHz (500 MHz to 1.0 GHz) are largely due to the counters 500 MHz direct counting ability (Fig. 1). At the band edges, fm tolerance is 30 MHz in the worst case. Even with this high degree of tolerance, sensitivity remains −15 dBm in the automatic mode and −20 dBm in the manual mode.

In the automatic mode of operation, the new counter will measure and display the lowest frequency cw signal within its sensitivity range. But in the manual mode, the operator can elect to search within any of ten frequency bands that cover the counter's full range. This is especially advantageous when attempting to find and measure several signals in a complicated spectrum or for applications where extremely low acquisition times are necessary. Acquisition time is slashed from 600 μs to 100 μs when measuring in the manual mode.

A diagnostic flow chart supplied with the counter guides the user through a self-check procedure that pinpoints failures to a modular level. Troubleshooting simply becomes a matter of observing the binary (zero or one) state of the three display digits that are activated when the internal test switch is thrown. The counter measures 500 MHz in its check mode (in contrast to the usual 10 MHz) in order to test the counting circuitry in a more demanding region.



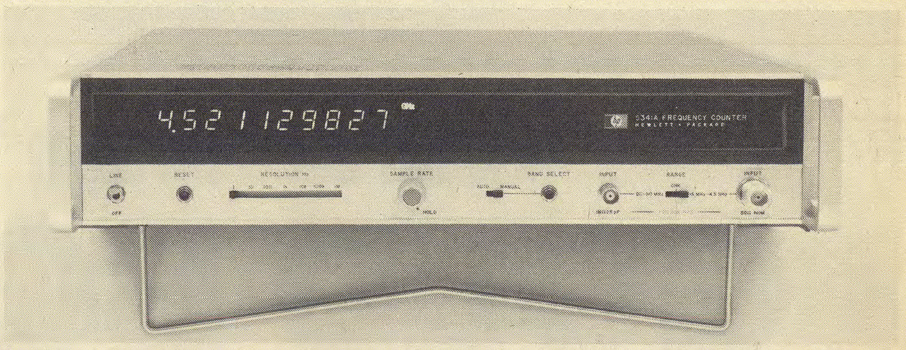
1. Thin-film switchable filters select the proper comb line for heterodyne mixing. The modular design permits a low cost 1.5 GHz option.

Ten LED digits provide 1 Hz resolution throughout the full range of the counter (Fig. 2). Input frequencies from 50 MHz to 4.5 GHz are channeled in through a 50-ohm type N connector, while inputs ranging from 10 Hz to 80 MHz meet an impedance of 1 MΩ shunted by 50 pF through a BNC port. Sensitivity of the low frequency input is 10 mV and the max input is 5 V peak-to-peak.

The crystal oscillator time base operates at a frequency of 10 MHz. Standard aging rate is less than 1×10^{-7} per month, but an optional time base (option 001) ages less than 5×10^{-10} per day after a 24 hour warm-up. Temperature stability is normally $\pm 1 \times 10^{-6}$ from 0 to 50°C, but option 001 improves this to 7 parts in 10^9.

Other features include an accuracy of ±1 count ± time base error; a front panel variable resolution control, calibrated at 1MHz, 100 kHz, 10 kHz, 1 kHz, 100 Hz, 10 Hz and 1 Hz; and a sample rate that is continuously adjustable from 40 ms to 10 sec and HOLD. Operating temperature range is 0 to 50°C. The counter weighs 10.5 kg (23 lb) and measures approximately 425 × 467 × 88.2 mm (16.75 × 13.25 × 3.5 in.). P&A: model 5341A (4.5 GHz): \$3,600: high stability time base (option 001): \$500; rear panel inputs (option 002): \$105; 1.5 GHz top range (option 003): less \$1,000; remote programming-digital output (option 001): \$390. **Hewlett-Packard, 1501 Page Mill Road, Palo Alto, CA 94304 (415) 493-1501.**

CIRCLE NO. 104



2. Ten LED digits can provide 1 Hz resolution for 4.5 GHz measurements.

Wideband Phase Modulator Works Directly On Carrier

Subharmonic modulators have many advantages, but suffer from excessive incidental am enhancement. This detrimental effect can be slashed in half if the output frequency is modulated directly.

MICROWAVE carrier frequencies are often phase modulated at a subharmonic level, then multiplied in a varactor multiplier to obtain the desired output. The advantages of doing this, when a large modulation index (±90 degrees) is needed, are quite clear. If the multiplication factor (n) is large, the modulation index required at the subharmonic is small and easy to obtain. Linearity is likely to be good, since only 1/n of the output phase deviation is actually needed in the modulator.

This approach has one serious disadvantage, however, and that is the undesirable effect that the varactor multiplier may have on the modulation spectrum. High-order multipliers are usually narrow-band devices with a variety of potential instability problems[1,2,3]. Incidental am enhancement, pm-to-am conversion and distortion of the modulator's baseband frequency response are often the results. These problems are particularly noticeable if the modulating frequency is several megahertz or higher.

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The modulator circuit shown in Fig. 1 eliminates these multiplier induced problems, since it operates directly at the desired output frequency.

When properly compensated, it is linear to within less than ±4% deviation from the best straight line over a 190-degree phase excursion. Incidental am is less than 10%. A sub-harmonic modulator might be just as linear, but incidental am could be at least two or three times as great due to pm-to-am conversion and negative resistance effects in the multiplier. Also, transmitting high frequency modulation through the multiplier can trigger relaxation oscillations that render the data useless.

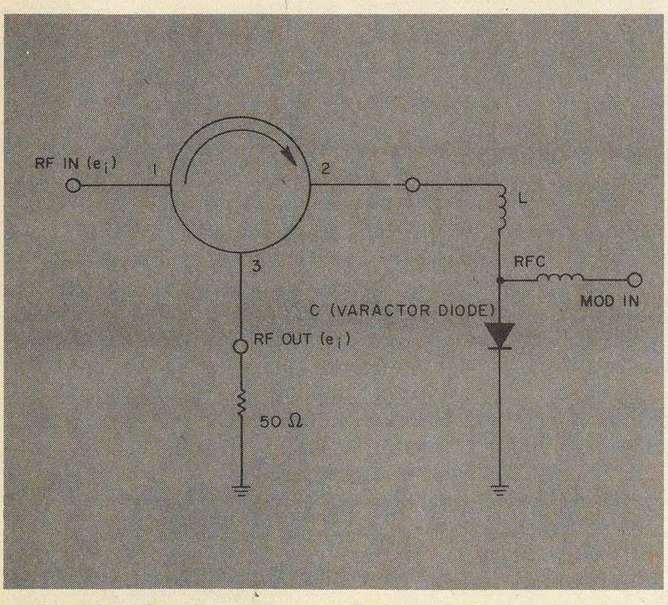
Here's how it works

The rf signal (e_i) enters port one of the circulator and travels to port two, which is terminated in a varactor-tuned series LC circuit. Assuming a purely reactive termination, all incident power is reflected to port three. Phase modulation is possible since the angle of the reflected voltage (e_r) varies with tuning of the diode. The relationship between e_r and e_i is given by the equation for reflection coefficient:

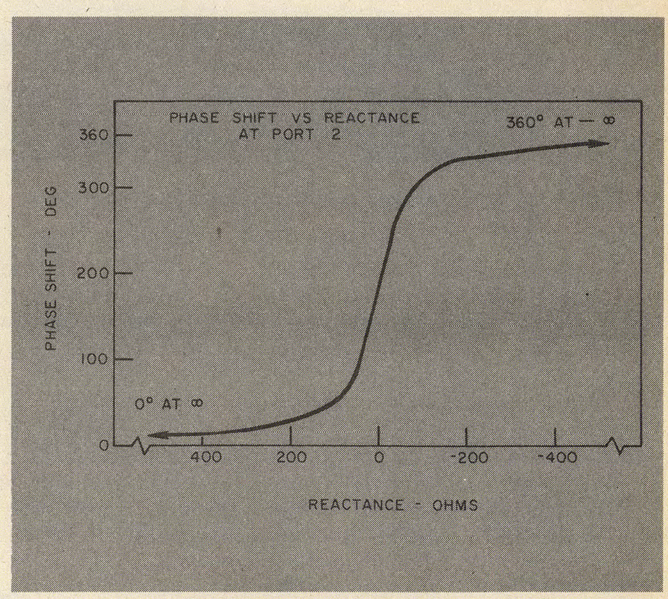
$$\frac{e_r}{e_i} = \frac{Z - Z_o}{Z + Z_o} \qquad (1)$$

Where: e_i is the incident voltage at port one,
e_r is the reflected voltage at port three,

(continued on p. 54)



1. **By directly modulating** the output frequency, this circuit circumvents many multiplier-induced problems.



2. **The theoretical phase relationship** of incident and reflected signals shows a linear portion of about ±100°.

Z_o is the characteristic impedance of the circulator and

Z is the impedance at port two.

If, as in the case under consideration, Z is purely reactive and Z_o is purely resistive, Eqn. (1) can be manipulated to give:

$$\frac{e_r}{e_i} = \frac{\sqrt{Z^2 + Z_o^2} \angle \tan^{-1}\left(-\frac{Z}{Z_o}\right)}{\sqrt{Z^2 + Z_o^2} \angle \tan^{-1}\left(\frac{Z}{Z_o}\right)} \quad (2)$$

which simplifies to:

$$\frac{e_r}{e_i} = 1 \angle -2\tan^{-1}\left(\frac{Z}{Z_o}\right) \quad (3)$$

Equation (3) indicates that the circuit in Fig. 1 can produce up to 360 degrees of phase modulation.

A parameter of prime interest in a phase modulator is linearity. Figure 2 is a plot of the calculated phase relationship between e_r and e_i as the reactance at port two varies from $+j\infty$ to $-j\infty$. Note that there is approximately a ± 100-degree region in which a linear change in reactance at port two causes a linear change in phase at port three.

Unfortunately, varactor diodes do not have a linear characteristic. More commonly, their reactance varies as the square root of the applied voltage. But adding a simple inductor (L_1) linking port two with ground (Fig. 3) provides reasonably good compensation.

Two diodes are used to simplify biasing (the top diode prevents L_1 from shorting the dc bias voltage to ground) and to obtain a push-pull effect that tends to prevent large rf signals from detuning the circuit.

L_2 and C series resonate when the modulation input voltage is -3 V. In this low-voltage region, the reactance of the diode changes quickly with voltage ($X_c \propto \sqrt{V}$); but as the voltage goes more negative, the reactance changes more slowly which causes a flat region in the phase versus voltage curve. When L_1 is introduced to the circuit and the varactor voltage is less than -3 V, the L_2C branch looks like a capacitor that begins to parallel resonate with L_1. The approach of parallel resonance forces the rate of change in reactance to increase which tends to cancel the effect of the more slowly changing varactor reactance.

The equation for the reactance at port 2 is

$$Z = \frac{(jX_2 - JX_c)\, jX_1}{jX_2 - jX_c + jX_1} \quad (4)$$

where:

X_1 = reactance of L_1

X_2 = reactance of L_2

X_c = reactance of C

In order to show X_c as a function of voltage and to reduce the labor involved in calculations, Eqn. (4) can be written as:

$$Z = \frac{\left(jX_2 - jX_2\sqrt{\frac{V}{K}}\right) jmX_2}{jX_2 - jX_2\sqrt{\frac{V}{K}} + jmX_2} \quad (5)$$

where

V = modulation voltage and

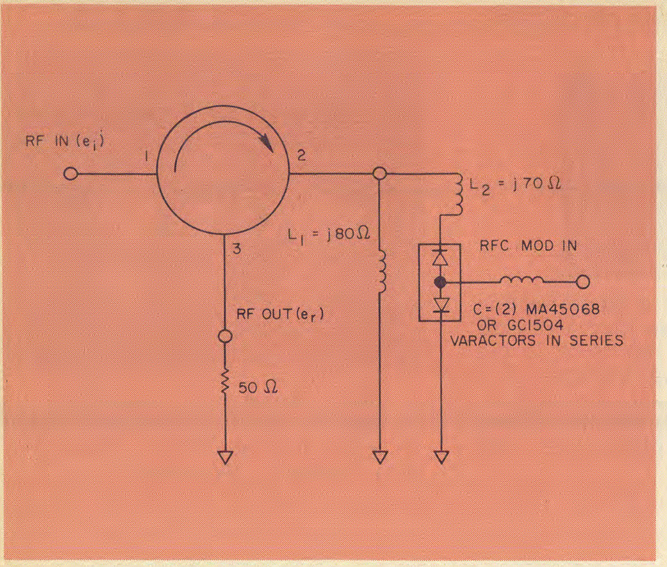
K = modulation voltage at which $jX_2 = -jX_c$

$m = X_1/X_2$

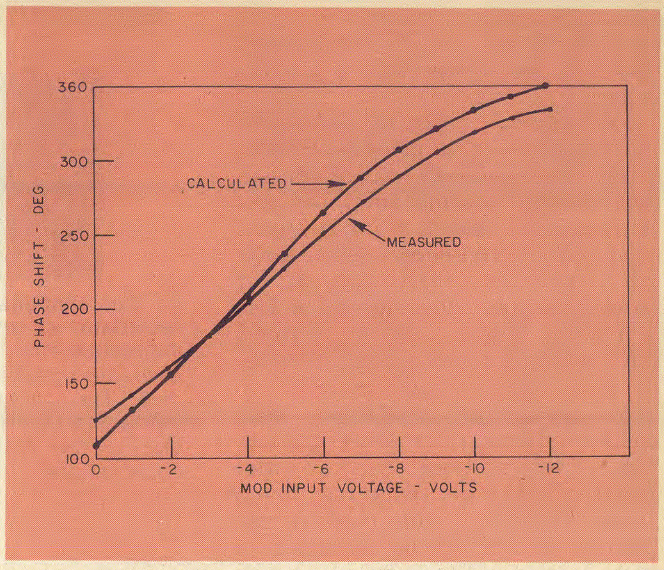
This reduces to:

$$Z = \frac{jX_2\left(1 - \sqrt{\frac{V}{K}}\right) m}{1 - \sqrt{\frac{V}{K}} + m} \quad (6)$$

The transfer function for the modulator is now obtained by substituting (6) into (3):



3. **L_1 is necessary** to compensate for the non-linear response of the varactor diode.



4. **A measured phase characteristic** shows that more than ±90 degrees of linear deviation is available.

$$\frac{e_r}{e_i} = -2 \tan^{-1} \frac{\left[\left(\frac{X_2}{Z_0}\right)\left(1 - \sqrt{\frac{V}{K}}\right) m\right]}{1 - \sqrt{\frac{V}{K}} + m} \qquad (7)$$

Experimentally it has been found that good linearity is obtained with the following circuit parameters:

$$X_2 = -j70$$
$$m = 1.143$$
$$\frac{X_2}{Z_0} = 1.40$$
$$K = 3$$

A plot of the calculated and measured performance obtained with these values is shown in Fig. 4.

High-Q components necessary

Earlier, it was assumed that the tuned LC circuit is purely reactive. In reality, of course, this is not the case. A small resistive component absorbs a portion of the incident power which results in in-

(continued on next page)

cidental amplitude modulation. The amount of am can be kept small, however, by using high-Q components.

For example, Fig. 5 illustrates the response of a modulator with a 450 Hz sawtooth waveform applied to the modulation input. As can be seen, a ±5.0 V input signal produces a ±100-degree phase deviation with less than ±0.5 dB of incidental am. The insertion loss is 1.5 dB. From −30° to +70°C, the modulation index remains within ±1 dB of the value at 23°C.

A single-stage modulator may not be sufficient for every application. Two sections can be cascaded to obtain twice as much phase deviation (the total deviation is the sum of the deviation in each section). Or, better linearity can be obtained by operating each section at, say, one-half as much deviation to avoid the most nonlinear region in the phase versus voltage curve.

A two-section modulator is shown in Fig. 6. The specifications (Table 1) for this design show the potential of a multi-stage approach. The maximum specified power level (20 mW) is based on stability rather than burnout considerations. At high-drive levels, the modulator may behave like a parametric amplifier and break into oscillation. Under the ideal conditions, satisfactory performance has been obtained at up to a 100 mW input, but this much drive is not recommended.

Linear frequency deviation possible

In addition to providing linear phase deviation, the modulator can also produce linear frequency deviation when used in the oscillator design depicted in Fig. 7. The circuit shown oscillates at approximately the center frequency of the bandpass filter, and exactly at the frequency at which the loop phase shift is an integral multiple of 2π radians.

All components in the oscillator should be wideband, with the exception of the filter, which should have a flat group delay versus frequency characteristic in the bandpass region. If this criteria is met, the oscillator's stability is determined primarily by the filter, and its deviation linearity primarily by the phase modulator. This is illustrated in the following analysis which is analagous to Kirchhoff's voltage law with phase corresponding to voltage, frequency to current and group delay to resistance.

Let Θ be the phase shift through each element in the loop. Since the total phase shift around the loop in an oscillator must be integral multiple of 2π radians[4] or 0, we may write:

$$\Theta_a + \Theta_c + \Theta_m + \Theta_f = 2\pi\, n \quad (8)$$

where:

Θ_a = the phase shift in the amplifier

Θ_c = the phase shift in the coupler.

Θ_m = the phase shift in the modulator

Θ_f = the phase shift in the filter

n = an integer

Taking the derivative of Eqn. (8) with respect to frequency (ω) gives:

$$\frac{d\Theta_a}{d\omega} + \frac{d\Theta_c}{d\omega} + \frac{d\Theta_m}{d\omega} + \frac{d\Theta_f}{d\omega} = 0 \quad (9)$$

Each of these terms is the group delay (τ) of one element in the oscillator. Hence, we may write:

$$\tau_a + \tau_c + \tau_m + \tau_f = 0 \quad (10)$$

The delay in the modulator is

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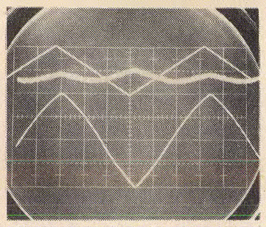
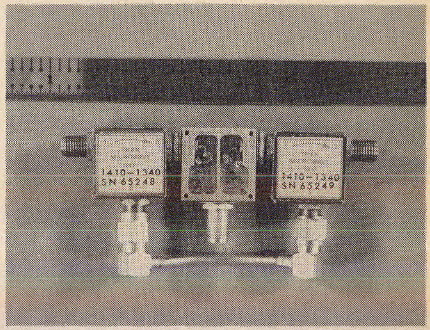
MODULATOR PERFORMANCE (ONE SECTION)



5. **This response of a single-section modulator** shows less than ±0.5 dB incidental am. The top trace is the sawtooth modulating voltage (5V/div). The center trace shows incidental am (1 dB/div) and the bottom one is the actual phase shift (50 degrees/div).

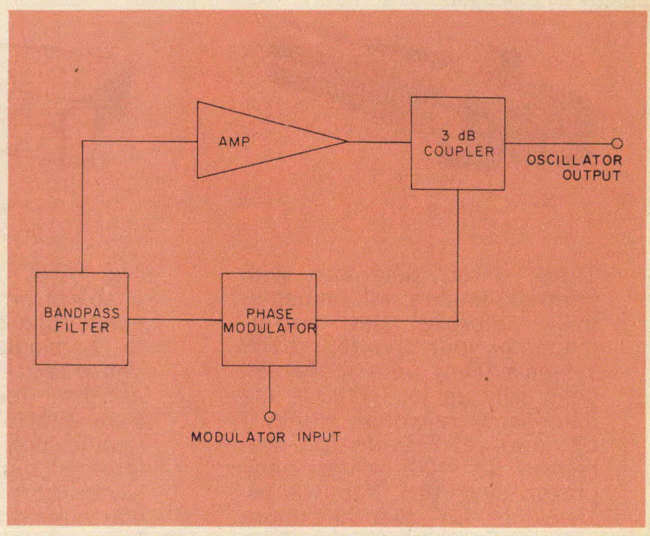
TWO SECTION MODULATOR



6. **This two-section modulator** is capable of ±180-degree modulation or ±90 degrees at less than 3% total harmonic distortion.

Table 1: Two Section Modular Specifications

Frequency	2.1-2.3 GHz
Modulation capability	±180° (typical)
Linearity	<3% harmonic distortion at ±90°
Modulation frequency response	dc to 55 MHz (upper 3 dB frequency)
Modulation input impedance	>1 MΩ shunted by 10 pF
Insertion loss	≤3 dB
Incidental am	± 0.5 dB at ± 90°
Modulation sensitivity	± 3V/100° (typical)
Input/output impedance	50 ohms
Maximum rf input	20 mW
Size	6.5 x 2.0 x 1.2 cm
Weight	71 grams



7. **Linear frequency deviation** is taken advantage of in this oscillator application.

assumed to be negligibly small and constant in the frequency region of interest. The latter condition permits us to write:[5]

$$\tau_m = -\frac{\Delta\Theta_m}{\Delta\omega} \qquad (11)$$

Substituting Eqn (11) into Eqn (10) and solving for $\Delta\omega$ results in:

$$\Delta\omega = \frac{\Delta\ \Theta_m}{\tau_a + \tau_c + \tau_f}$$

This equation illustrates that if a phase change $\Delta\Theta_m$ is forced into the loop by the modulator, the oscillator frequency must change by $\Delta\omega$ in order to generate an equal and opposite phase change. Equation 12 holds true when the time interval for $\Delta\Theta_m$ is large compared to $(\tau_a + \tau_c + \tau_f)$.

If the group delay in each loop is constant, the linearity of the frequency deviation will be equal to the linearity of phase deviation. In some cases where the loop group delay is not constant, it may be possible to modify the modulator phase characteristic to compensate for phase nonlinearities in the loop.

The oscillator in Fig. 7 offers several practical features. Since each circuit element has a 50-ohm input-output impedance, each may be treated as a separate module and interconnected with 50-ohm cable. This is advantageous since it simplifies open loop measurements on a network analyzer.

An apparent disadvantage is that the output signal can abruptly jump to a frequency above or below the desired frequency by $\frac{2\pi n}{\tau}$, where n is an integer and τ is the total loop group delay. This problem can be avoided, however, if the bandwidth of the loop filter is narrow enough that the loop gain at the unwanted frequency is less than one. ••

Acknowledgement

The author would like to thank Dave Hepler for computer programs and interesting discussions regarding the modulator.

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Test your retention

1. What simple component can be put in parallel with a varactor-inductor series circuit to compensate for the varactor's non-linear response?
2. How would a modulator be used in an oscillator circuit?

Transistor Mixer Boosts Upconversion Gain

A transistor-based design for a uhf broadcast upconverter can provide five times more gain than a similar diode design. Here's a comparison of both circuits, each built inside a rat race hybrid.

RF power transistors operated in the so-called harmonic mode can be the basis for some interesting high-power upconverters. An overlay transistor operating in this mode generates harmonics or acts as a parametric upconverter due to a non-linear collector-base junction capacitance—an effect that is very similar to the action of a varactor diode. In addition, the transistor provides power gain at the fundamental input frequency giving the effect of an amplifier in parallel with a varactor circuit.

The advantages of using transistors instead of varactors can best be demonstrated by comparing similar circuits designed with each device. A good design example is a family of parametric upconverters designed primarily to upconvert a 38.9 MHz video i-f to a vhf, uhf or X-band broadcast frequency. The circuits to be compared are push-pull designs built inside a 6/4-wavelength "rat race" hybrid ring. Both microstrip designs are fabricated on copper-clad Rexolene 1422.

The basic parametric upconverter circuit using varactor diodes is shown in Fig. 1. An input signal (ω_s) enters through port one, where a small transformer (Fig. 2) wound on a Siemens' two-hole ferrite-core (Type 80 K 1) matches the 50-ohm impedance to the impedance of each varactor circuit, which can essentially be represented by R3 and R4 (each 15 ohms). The two diodes are push-pull driven through the input transformer, as well as by the pump power (Ω_p), which is introduced through port two on the hybrid ring. Trimmer capacitors (C1 and C2) match the diodes to the hybrid ring.

Upper ($\Omega_p + \omega_s$) and lower ($\Omega_p - \omega_s$) sideband signals are available at port three. One of these can be selected by an appropriate filter. For example, a three-section coaxial bandpass filter can provide more than 60 dB suppression to unwanted frequencies. With a 0.5-watt input, this varactor circuit produces about 1.5 watts in the lower sideband between 470 and 960 MHz. Pump power is in the range of six to seven watts.

Instead of specifying efficiency, it's more useful to deal with the conversion gain of an upconverter. This gain is defined as the ratio of upconverted signal power to the input signal power. Therefore, this

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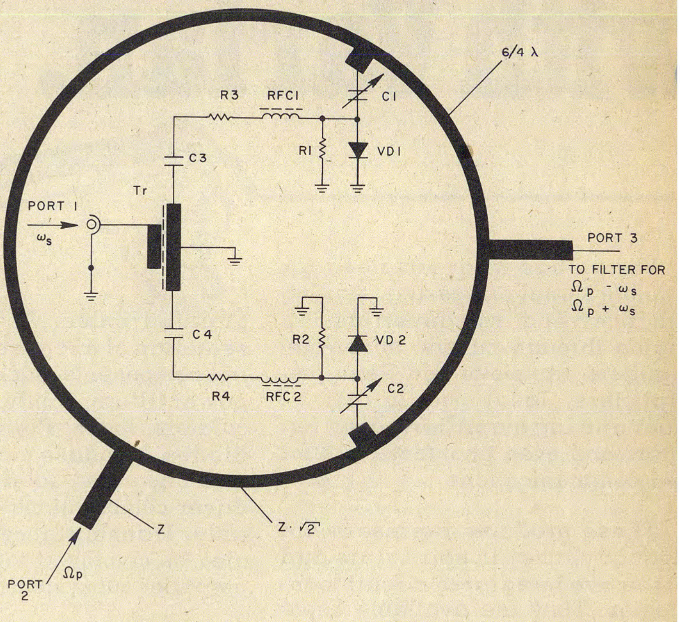
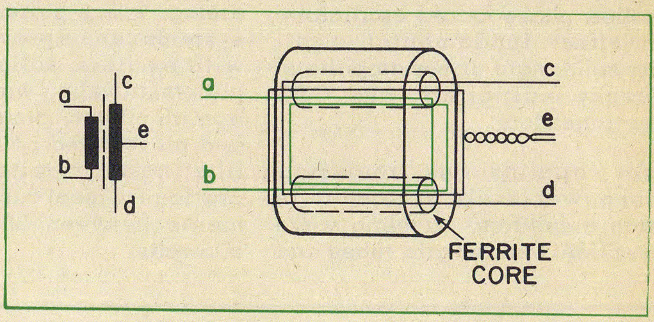


Component Values

C 4.7: nF bypass capacitor.
C 1, C 2: 1 to 10 pF miniature air-dielectric trimmer capacitor (Tronser or equivalent type).
C 3, C 4: 10 nF, 30 Vdc, miniature disc ceramic capacitor (Siemens or equivalent type).
C 5, C 7, L 1 and C 6, C 8, L 2:
Low-pass filter (π-section), values depending on input signal frequency range; input/output impedance approximately 15 ohms.
R 1: 1 kΩ, 0.5 watt, carbon resistor.
R 2: 500-ohm, 1 watt, linear

1. A simple 6/4 wavelength hybrid ring surrounds this varactor upconverter. An upconversion gain of 3 is possible at uhf.



2. This double-hole ferrite-core transformer matches the 50-ohm input impedance to the 15-ohm varactor circuits.

varactor circuit has a conversion gain of:

$$g_v = \frac{1.5}{0.5} = 3$$

C-B junction acts like a varactor

A comparison of varactor and transistor performance (Fig. 3) reveals that the nonlinear capacitance (C_{bc}) characteristic of a transistor's collector-base junction is very similar to the variation of a varactor diode's capacitance with changes in reverse voltage. A common-base transistor circuit (Fig. 4) corresponds to the varactor circuit described above, since the collector-base capacitance of the transistor is connected in shunt with the signal and pump inputs as well as paralleling the output circuit.

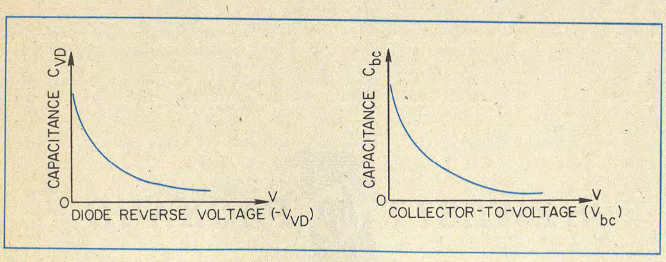
Using the transistors in a common-base configuration also provides excellent stability, good isolation and a high cutoff frequency. In practice, operation is limited by two distinct cutoff frequencies. One is the common-base current gain cutoff, $f_{\alpha b}$, which is the frequency at which h_{fb} has decreased to a value of 3 dB below h_{fbo}. This is normally much higher than $f_{\alpha e}$, the common emitter current gain cutoff, for any given transistor. The second cutoff factor, the cutoff frequency of the collector-base junction, is limited only by the collector series resistance. With the RCA 2N4012 transistors used in this design, substantial gain can be realized for input frequencies up to 450 MHz.

An added bonus is accrued when using the transistors in a common-base configuration. The grounded base acts like a shield between input and output circuits effectively preventing any part of the pump power or upconverted signal power from feeding back to the input circuit.

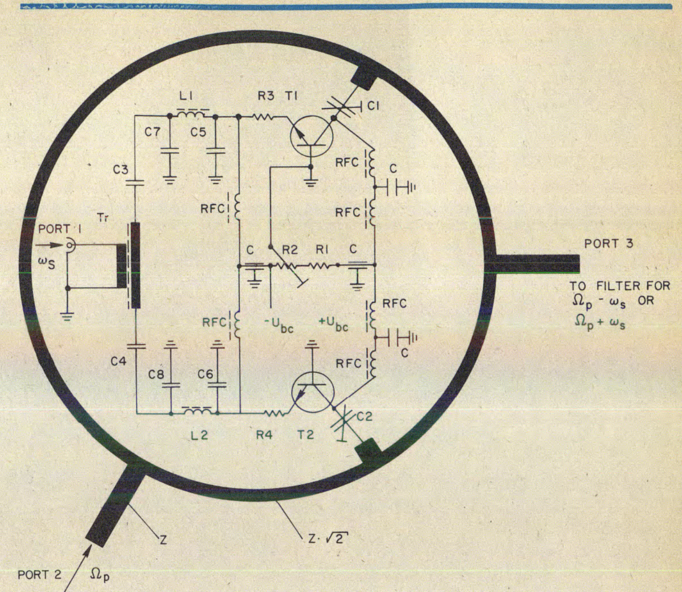
Many of the components used in the varactor upconverter can also be used in the transistor version. Input signals, introduced at port one, pass through a similar ferrite-core transformer, since the actual transistor input impedance is negligible compared to R3 and R4. A π-section low-pass filter supplements the input circuit of this design for improved impedance matching. With the cascaded transformer-filter arrangement, the transistor circuit has an input bandwidth of 30 to 70 MHz. The collector circuits are capacitively coupled to the hybrid ring.

5:1 improvement in gain

With six or seven watts of pump power applied to port two, a 100 mW input signal results in an output power of greater than 1.5 W in the lower sideband for uhf television channels in the frequency range of 470 to 960 MHz. Under comparable conditions, the transistor circuit's upconvertion gain is 15—five times that of the varactor design. This indicates that the transistor circuit needs far less input power.



3. The collector-base capacitance of an overlay transistor is similar to the capacitance of a varactor diode. This feature allows frequency translation with added power gain.



Component Values

C 1, C 2: 1 to 10 pF air-dielectric miniature trimmer capacitor (Tronser or equivalent type).
C 3, C 4: 10 nF, 30 Vdc, miniature disc ceramic capacitor (Siemens or equivalent type).
R 1, R 2: 50 K 0.25-watt, bias resistor; carbon type with axial lead wires.
R 3, R 4: 15-ohm, 0.5-watt, input matching resistor; carbon type with axial lead wires.
RFC 1, RFC 2: Rf choke, consisting of a piece of 0.4 mm diam. wire, one-quarter of a wavelength (of the pump frequency) long, turned to a simple coil of 6 to 8 mm inner diameter.
VD 1, VD 2: Varactor diodes. Microwave Associates MA 4060B or equivalent.
Tr: Push-pull transformer (Fig. 2).
Ports 1, 2 and 3: TNC or BNC flange-type connections.
trimmer resistor or potentiometer (Dralowid or equivalent type) for base-bias voltage setting.
R 3, R 4: 15 ohm, 0.5 watt, emitter input matching resistor; carbon resistor with axial lead wires.
RFC: Rf choke, 6µH vhf choke with ferrite core and shrunk sleeve (Siemens or equivalent type).
T 1, T 2: Overlay transistor, RCA 2N4012, 2N5016 or equivalent.
Tr: Push-pull transformer (Fig. 2).
Ports 1, 2 and 3: TNC and BNC flange type connectors.

4. Transistor upconverters feature a high upconversion gain. A common-base circuit is used to take advantage of a varactor-like collector-base capacitance.

Both of these designs have been used in Germany as television broadcast upconverters. They are very simple to build. The microstrip "rat race" ring is not very critical with respect to its wavelength, and is made with a 10% tolerance. Strip width (ring impedance) can be calculated or taken directly from design curves. The substrate is fastened to a sheet of metal, which holds the transistors and acts as a simple heat sink. Flanged TNC or BNC connectors are used on ports one, two and three. The finished unit may simply be mounted in a metal box for complete shielding. ●●

Test your Retention

1. **Define conversion gain.**
2. **How long is a rat-race?**
3. **How should an overlay transistor be mounted to obtain a varactor-like response?**

Transplexing SAW Filters For ECM, *Part II*

Miniature amplifiers can be effectively combined with SAW filters to form filterbanks with contiguous passbands. Advantages include insertion gain and a dynamic range of 45 dB.

This is the second of a two-part article on surface-acoustic wave filters and their use in multiplexed filterbanks. Part I appeared last month.

THE performance of a filter array with contiguous passbands is citically dependent on the manner in which the individual filter elements are interconnected. Usually it is not possible to simply parallel the filter elements and achieve the response of the isolated filters. The effects of channel interaction and source loading must be considered.

One common method of interconnecting filter elements[14] is shown in Fig. 1. The individual filter elements must be designed from a singly-terminated prototype since the input admittance of adjacent channels combines to form an equivalent voltage source at the terminals of a given filter. Also, the first element of each channel must have the characteristics of a series resonant circuit in order that its input conductance be zero for out-of-band frequencies and increase to a value equal to the source conductance for in-band frequencies. Since the power delivered to a given channel is proportional to the input conductance, the transmission characteristics of that channel are also proportional to the conductance. Thus when the specially-designed filters comprising the the multiplexer are paralleled, the real part of the total input admittance, Y_T, is relatively constant across the multiplexer bandwidth with a value equal to G_s. The imaginary part, B_T, is approximately zero.

There are many available techniques for multiplexing including the use of circulators to reduce adjacent channel interaction. However, all these techniques require that the input reflection coefficient of a filter element be zero across its passband. With surface wave acoustic filters, this is generally not the case due to the parasitic transducer capacitance.

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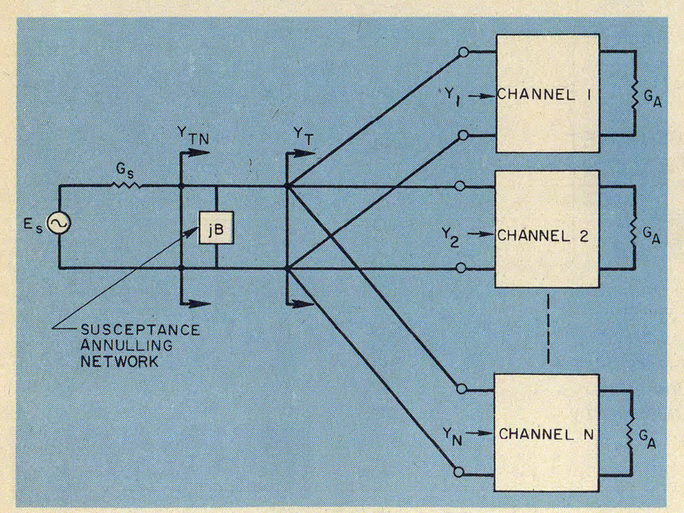
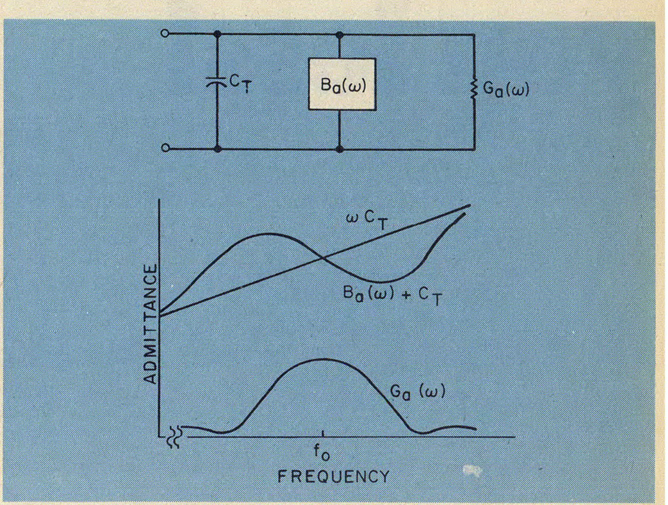
Active "transplexing"

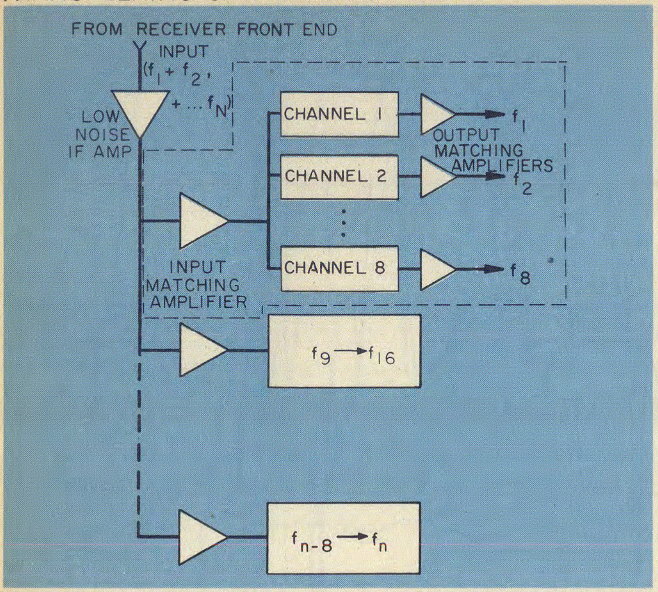
Using the "crossed-field" model, Fig. 2, the input admittance of the interdigital transducer may be characterized by a circuit consisting of the total electrode capacitance in parallel with a radiation admittance which represents acoustic-wave excitation. The input power must be efficiently delivered to the real part of this radiation admittance. Note that $B_a(\omega)$ is zero and G_a is a maximum at band center, f_o, and that if C_T were zero, the input admittance would be similar in form to the multiplexer in Fig. 1. However, for transducers on a substrate with a low electromechanical coupling constant k such as quartz, the ratio of $\omega_o C_T$ to $G_a\ \omega_o$, which is defined as the radiation Q, may be as high as 100. Fortunately for narrow-bandwidth filters, Q_r decreases due to the increase in the number N of interdigital periods required by,



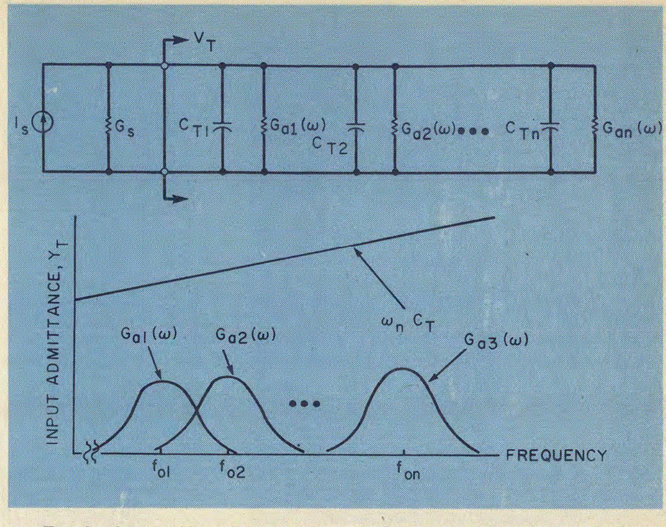
1. **For interconnecting filter elements,** the first element of each channel must present a series resonant circuit to reject out-of-band frequencies.



2. **Cross-field model oi an interdigital transducer.** Input power must be efficiently delivered to the real part of the radiation admittance, Ga.



3. **In this n-channel filter array** active transplexing makes use of input/output matching amplifiers to drive up to eight SAW filters.



4. **Equivalent circuit of an n-channel filterbank** when driven by a single matching amplifier.

$$Q_r = \frac{\pi}{4}\frac{1}{k^2N} \qquad (2)$$

However, Q_r may still be as large as 10-20 for filters with a bandwidth as small as 1%. Therefore, due to the presence of C_T, active impedance transforming, as opposed to passive networks, becomes an attractive method for interfacing contiguous acoustic filters to form a multiplexer. In active "transplexing," (transducing-multiplexing) a single matching amplifier drives a set of filters. This forms a broadband magnitude match with the total input capacitance C_T. Typically, eight filters are driven (as shown within the dotted lines of Fig. 3) but the number of channels is dependent on the maximum input Q which the amplifier can efficiently drive. A simplified input equivalent circuit for n channels in parallel driven by a single matching amplifier is given in Fig. 4. The radiation susceptance, B_a, is neglected since it is much smaller that ωC_T. One advantage of high-transducer susceptance is that it effectively masks the radiation admittance and thus minimizes interchannel interaction. Assuming that the filters have equal transducer capacitance and radiation Q, the total input Q is approximately n times that of a single channel. If the source conductance is adjusted equally to the load susceptance, the power delivered to the radiation conductance of a given channel is down by a factor of n from a channel driven by a separate matching amplifier.

An output matching amplifier at each channel also serves as an active admittance transformer, matching the output transducer to the 50-ohm load. It also provides gain to compensate for any remaining losses. The gain of the output-matching amplifier can easily be made variable, and, therefore, provides a convenient means of equalizing the signal output level of the filter array.

Figure 3 is an n channel filter array consisting of n acoustic filters and output matching amplifiers. An input matching amplifier is required for each set of eight filters. The input admittance of the individual input-matching amplifiers is designed to form a broadband power divider over the multiplexer bandwidth. Additional channels can be added to the multiplexer in sets of eight by simply adding another subsection in parallel. Due to the reverse isolation of the input matching amplifiers and their broadband input admittance, channel-to-channel interaction is minimized.

Selectivity determines dynamic range

One possible limitation to active "transplexing" is due to the nonlinear nature of active devices at high-power levels. This establishes the upper-power level which can be applied to the filterbank. However, if the matching amplifiers are properly designed, this power limit does not restrict the system's dynamic range. The available dynamic range of a SAW filterbank with matching amplifiers is typically 90 dB—bounded by the amplifier one dB compression point at the upper end and noise level at the lower end. However, due to the finite selectivity of acoustic filters (typically 56-60 dB), the practical limitation is the filter selectivity and not non-linear effects.

The level of intermodulation products is a convenient measure of linearity. It is also an important consideration in its own right with a frequency multiplexer. If two signals mix in the input-matching amplifier thereby generating a third, this spurious signal could be interpreted as a "real" signal and result in a false alarm.

The third order ($f_1 \pm 2f_2$) intermodulation product is usually the largest spurious response, can be expressed as:

$$P_{IM} = P_1 + 2P_2 - 2P_{sat} - 18 \quad (3)$$

where

P_1 = power level (dBm) at f_1
P_2 = power level (dBm) at f_2
P_{sat} = power level (dBm) for which the gain is reduced 1 dB

Actively "transplexed" filterbank

An example of a four-channel filterbank using one input and four output amplifiers is shown in Fig. 5. It uses frequency contiguous quartz filters with 5 MHz bandwidths at 200 MHz. The frequency response of the filterbank, Fig. 6, shows the nominal insertion gain for each of the four channels to be 13 dB. The relative gain distribution in each input channel is 4 dB from matching and 21 dB from the input amplifier. On the output, 26 dB is achieved with the output amplifiers and 6 dB from matching. Thus, a total of 10 dB reduc-

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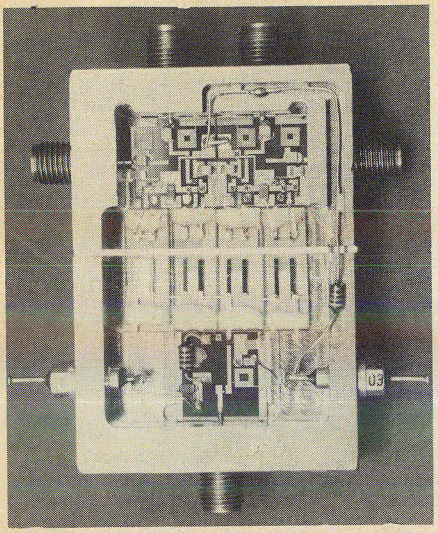
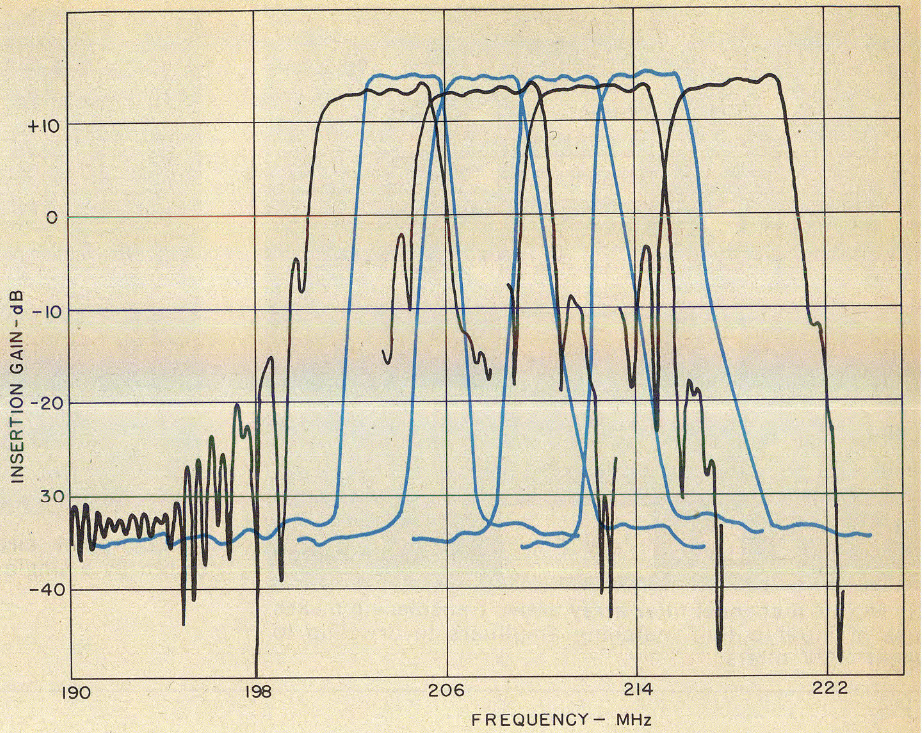


5. **Four-channel filterbank** is only 2.2 in. × 1.5 in. × 0.5 in. It uses SAW filters with 5-MHz bandwidth at 200 MHz.



6. **Frequency response of filterbank** (Fig. 5) shows a total insertion gain of 10 dB from the matching input/output amplifiers. "Electrically apodized" filters increase dynamic range as shown in color tint.

tion in insertion loss for each channel is achieved due to magnitude matching alone. The wideband (100-600 MHz) frequency response of channel one of the filterbank is plotted in Fig. 7. Dynamic range is near 30 dB. By employing new filter synthesis techniques which suppress diffraction effects and thereby reduces sidelobe levels, a dynamic range of greater than 50 dB with shape factor of 2:1 (45 dB to 3 dB frequency ratio) are possible up to and above 800 MHz. The predicted frequency performance of a multiplexer using the "electrically apodized" filter is shown in Fig. 6 (color tint). Triple transit leakage on this design is in excess of 60 dB. For most filter bank applications, a −40 dB intermodulation level is sufficient. However, if it is desired to reduce intermodulation products or raise the upper power limit, the input-matching amplifier can be modified to do this. One of the critical design tradeoffs is the amount of gain provided before the acoustic filters as opposed to after the filters. For minimum noise figure, it is desirable to put as much gain as possible before the filters. On the other hand, for minimum distortion, the gain should be placed after the filters. With the four-channel multiplexer, Fig. 5, the gain was proportioned equally between the input and output matching amplifiers. However, in most applications, sufficient system gain is provided before the filterbank to establish the system noise figure and, consequently, the filterbank has little effect on the system noise figure. Typical specifications for an n-channel filterband module are given in Table 1.

Cross-channel interference

An important consideration in the design of channelized filters is the effect of cross-channel interference on the dynamic range. Due to the finite slope of the filter skirts, a practical limit is imposed on the input dynamic range. In fact, the filter skirts are the single most important parameter for determining the dynamic range.

A "threshold" power level must be set for the purpose of deciding on the presence or absence of a signal. It must be set such that with minimum input power, P_{min} is at least 3 dB above the threshold. Then in the case of maximum

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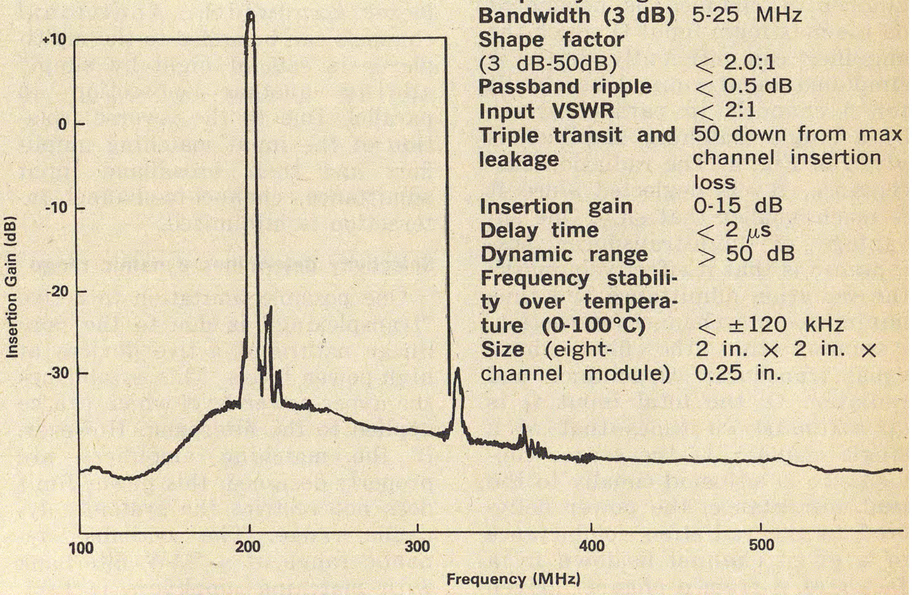


7. **Wideband frequency response** of a single channel of the filterbank (Fig. 5) shows dynamic range of over 30 dB.

Table 1 Filterbank characteristics

Frequency range	100-500 MHz
Bandwidth (3 dB)	5-25 MHz
Shape factor (3 dB-50dB)	< 2.0:1
Passband ripple	< 0.5 dB
Input VSWR	< 2:1
Triple transit and leakage	50 down from max channel insertion loss
Insertion gain	0-15 dB
Delay time	< 2 μs
Dynamic range	> 50 dB
Frequency stability over temperature (0-100°C)	< ±120 kHz
Size (eight-channel module)	2 in. × 2 in. × 0.25 in.

input power, the threshold will be exceeded in more than one channel, over a frequency band in the region of crossover. The frequency range in which two channels are triggered is given by

$$\Delta f_{max} = \Delta f \frac{\eta - 1}{\Delta \alpha} \Delta P_{in} \quad (4)$$

Δf = crossover bandwidth of the filters = 3 dB bandwidth

η = filter shape factor (defined as a bandwidth ratio between two attenuation points)

$\Delta \alpha$ = the difference in dB between the two attenuation points used to define η

ΔP_{in} = input dynamic range = $P_{max} - P_{min}$

For example, with η = 1.6 between the 54 dB and 3B points and ΔP_{in} = 40 dB.

$$\frac{\Delta f_{max}}{\Delta f} = 57\%$$

Therefore, in this example, two adjacent channels will be simultaneously triggered by a signal level P_{max} over a 57% baindwidth in the region of crossover.

In order to eliminate any such ambiguities, a double-detection scheme[1] could be used. This increases the effective filter selectivity, without exhibiting "ringing" for narrow pulses which would occur for the equivalent highly selective filter. The major drawback to this technique is additional complexity (therefore cost) incurred by using extra detectors and compartor circuits for each channel.

Quasi-monolithic filter bank

Significant cost reductions can be realized by using monolithic amplifiers in the actively "transplexed" filter bank. The monolithic amplifier, Fig. 8, is a (0.100 in. × 0.100 in.) chip fabricated by various epitaxial diffusions in N-type silicon. These chips are so designed to permit thermal compression wirebonding combinations of resistors, transistors, etc., into any configuration. Once the design is firm, a final overlay metallization may be added, and the amplifier design becomes monolithic. The f_T of the transistors used in the hybrid amplifiers is approximately 2 GHz; whereas, values of f_T near 650 MHz have been measured for the monolithic-bipolar transistors. Monolithic amplifier chips with transistor f_T as high as 2 GHz are under development. These amplifiers would be mounted on the quartz substrates and only the interconnects between filters and amplifiers would be thermal-compression wirebonds. Holding costs to a minimum dictate the use of a flatpack with the output connectors replaced by hermetic tabs.

The ultimate in cost-effective designs would be the fabrication of filters and amplifiers on one silicon substrate. Since silicon is not piezoelectric, ZnO transducers would be required to launch surface waves. However, the design would truly be monolithic since metallizations for filters and amplifiers would be deposited simultaneously. One serious problem with this approach is maintaining input to output electromagnetic isolation on silicon. Measurements indicate greater than 50 dB at this time.

Next: SAW filters as noise jammers

The use of surface-acoustic filters to synthesize stable discrete frequencies in the uhf frequency range was recently described by the Air Force Cambridge Research Labs[15]. The synthesizer is composed of a clock-controlled pulse generator which drives in parallel 21 frequency-contiguous SAW filters. The clock-pulse frequency and the filter design are configured in order to produce 21 continuously available cw output frequencies in the 500 MHz range separated by approximately 5 MHz.

The SAW noise jammer described in this article makes use of the same actively-transplexed filters for transmit and receive. Thus, the filters are deployed as frequency multiplexers during receive and during transmit the filters become noise generators. (In conventional noise jammer applications, the same filters are used for transmit and receive; however, during transmit a broadband noise source is inputed to the filters).

To understand this concept, it is helpful to review the operation of SAW oscillators. A basic SAW oscillator, Fig. 9, consists of an amplifier with a SAW filter (delay line) and a feedback element[16]. The advantages of a SAW oscillator are:

- Frequency is determined by the filter-tranducer pattern, not quartz crystal dimensions.
- Fundamental oscillators are possible at frequencies over 3 GHz (crystal oscillators have a 200-MHz maximum) using electron-beam fabrication.
- Fm capability exceeds 1%.
- Its excellent short-term frequency and temperature stability is close to that of AT-cut quartz cystals.
- It's low cost, simple and rugged.

Providing there is net-loop gain, oscillation will take place at a frequency given by:

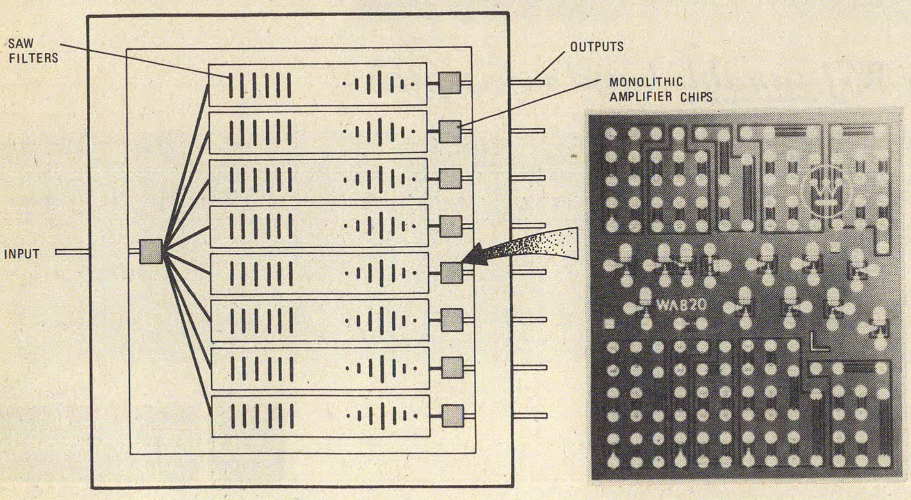
$$\frac{2 \pi f \ell}{V_s} + \phi_L = 2 n \pi \quad (5)$$

ℓ = center-to-center separation between input and output transducers

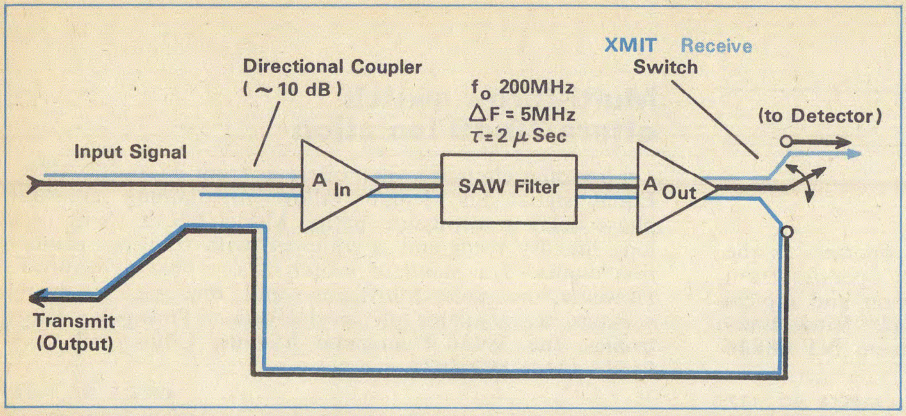
V_s = acoustic velocity

ϕ_L = loop phase shift

In an acoustic filterbank, by switching in a return line, each channel can be converted into an oscillator upon command via a PIN diode SPST switch. The frequencies of oscillation are determined by the center frequency, bandwidth and time delay of the acoustic filter. For example, a filter at 200 MHz with 5 MHz bandwidth and 2 μsec delay will oscillate at approximately 10 frequencies separated by 500 kHz. If the transmit-receive switch is opened and closed (during transmit cycle) at some random rate of three or four times the loop delay, all the frequencies within that filter passband are transmitted with an am modula-



8. **This monolithic amplifier** measures 0.1 × 0.1 in. sq. and is fabricated by epitaxial diffusions in N-type silicon.



9. **SAW oscillator** consists of an amplifier and SAW filter (delay line) with a feedback loop.

tion component. The SAW filter bandpass is designed for maximally flat response in order to achieve equal amplitude signals. Since the fm modulation rate is limited by the acoustic transmit time of the filter (2 μs which implies 500 kHz rate), the fm is accomplished by frequency modulating the upconverter LO. The combination of the am and fm modulation is used to simulate a noise transmission.

The filterbank could also be used as an "fm noise jammer" by switching in a return line (as opposed to cycling transmit-receive switch). After several circulations (depending on loop gain and filter design), only one frequency will remain, i.e., the loop gain for the other spectral lines will be less than unity. For this application, the acoustic bandpass filter must be designed to have lowest insertion loss at desired frequency of oscillation. As before, the modulation will be accomplished by frequency modulating the upconverter LO.

A n-channel actively "transplexed" transmit-receiver filterbank, Fig. 10, would sum all transmit signals by a resistive combiner and input to an upconverter. ●●

References

(continued from last month)

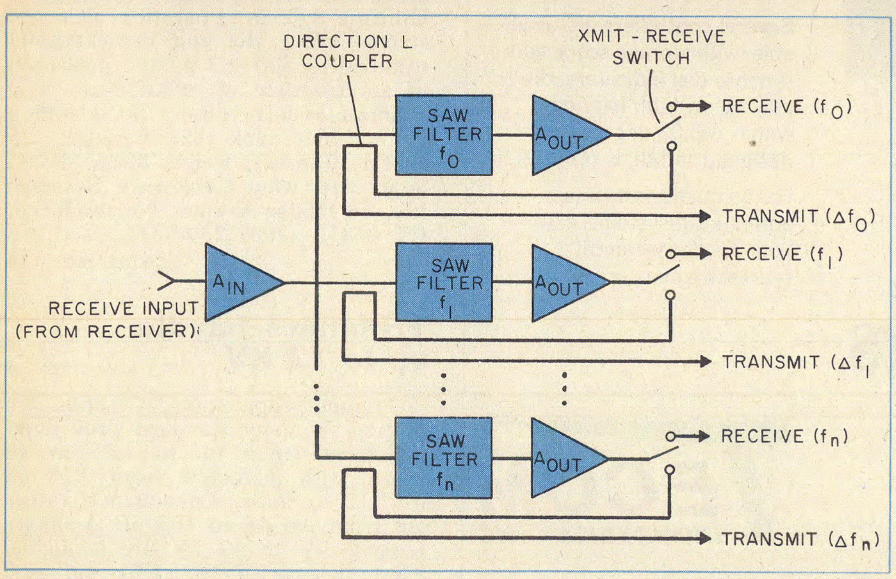
15. G. L. Matthaei, L. Young, E. M. T. Jones, *Microwave Filters Impedance, Matching Networks and Coupling Structures,* L. McGraw-Hill, New York (1964).

16. A. J. Budreau, Dr. P. H. Carr and Dr. K. R. Laker, "Frequency Synthesizer Using Acoustic Surface Wave Filters," *Microwave Journal,* pp. 65-68, (March, 1974).

17. M. F. Lewis, "Surface Acoustic Oscillators," *Microwave Systems News,* pp. 29-31, (December-January, 1974).

Test your retention

1. What are the three main functions of active "transplexing?"
2. What restricts the dynamic range of an actively transplexed filterbank?
3. For which other application is active transplexing uniquely suited?



10. **Proposed fm noise jammer** using SAW filterbank could automatically detect threat frequencies and transmit am and fm modulated noise signals back to the threat source.

new products

SWITCHES

Switch/driver costs $55

Rf/i-f diode switch/driver combination operates in the 3-300 MHz range. The SPDT unit which operates from a +24 V supply offers 60 dB min isolation and typical insertion loss of 0.75 dB. P&A: $55; stock. **Vitek Electronics, Ins., 200 Wood Avenue, Middlesex, NJ 08846 (201) 469-9400.**

CIRCLE NO. 117

Multi-throw switch offers 70 dB isolation

Fifty-ohm circular switch operates from dc to 700 MHz. Equal signal line lengths allow maintaining constant phase delay transmission paths. Model CSAZ-IX8B uses long life dry reeds and is equipped with transient protection diodes. The standard switch is one-pole-eight-throw. However, one-pole four-throw and one-pole six-throw versions are available on special order. **Trompeter Electronics, Inc., 8936 Comanche Avenue, Chatsworth, CA 91311 (213) 882-1020.**

CIRCLE NO. 116

PASSIVE COMPONENTS

Tiny circulator covers 7-17 GHz

Circulator/isolator, model T7S83T, covers 7-17 GHz and offers 16 dB isolation with an 83% bandwidth. VSWR is 1.45:1 while insertion loss is 0.75 dB. Temperature range is −54°C to +125°C and size is 0.78 × 0.63 × 0.56 in. **Teledyne Microwave, 1290 Terra Bella Avenue, Mountain View, CA 94040 (415) 968-2211.**

CIRCLE NO. 137

Calibration delay line trimmed to 500 ±3 ns

Model 83500 is a 500 ns calibration delay line for oscilloscopes, altimeters, radar and similar equipment. Utilizing 1/2 in. Foamflex dielectric coaxial cable, the unit is electrically trimmed to 500 ±3 ns and measured to an accuracy of ±0.020 ns. The calibration delay line is fitted with a 19 in. relay rack that measures 8.5 x 21 x 19 in. and weighs 80 lbs. P&A: $450; 6-8 wks. **Cablewave Systems, Inc., 60 Dodge Avenue, North Haven, CT 06473 (203) 239-3311.**

CIRCLE NO. 138

Trimmers handle up to 12 KV

Trimmer capacitors, VC series, are offered in many standard sizes ranging from 1.625 in. to 4.25 in. in length with diameters from 0.25 in. to 1.12 in. max. Capacitance values are from 0.6 pF to 100 pF. Voltages from 1 kV to 12 kV are available. P&A: in qtys of 100, $30; 4-6 wks. **Polyflon Corporation, 35 River Street, New Rochelle, NY 10801 (914) 636-7222.**

CIRCLE NO. 139