An Ultra-Stable Precision Demodulator for the Television Broadcaster

A 24 MHz Nyquist SAW Filter for the 1450 Demodulator

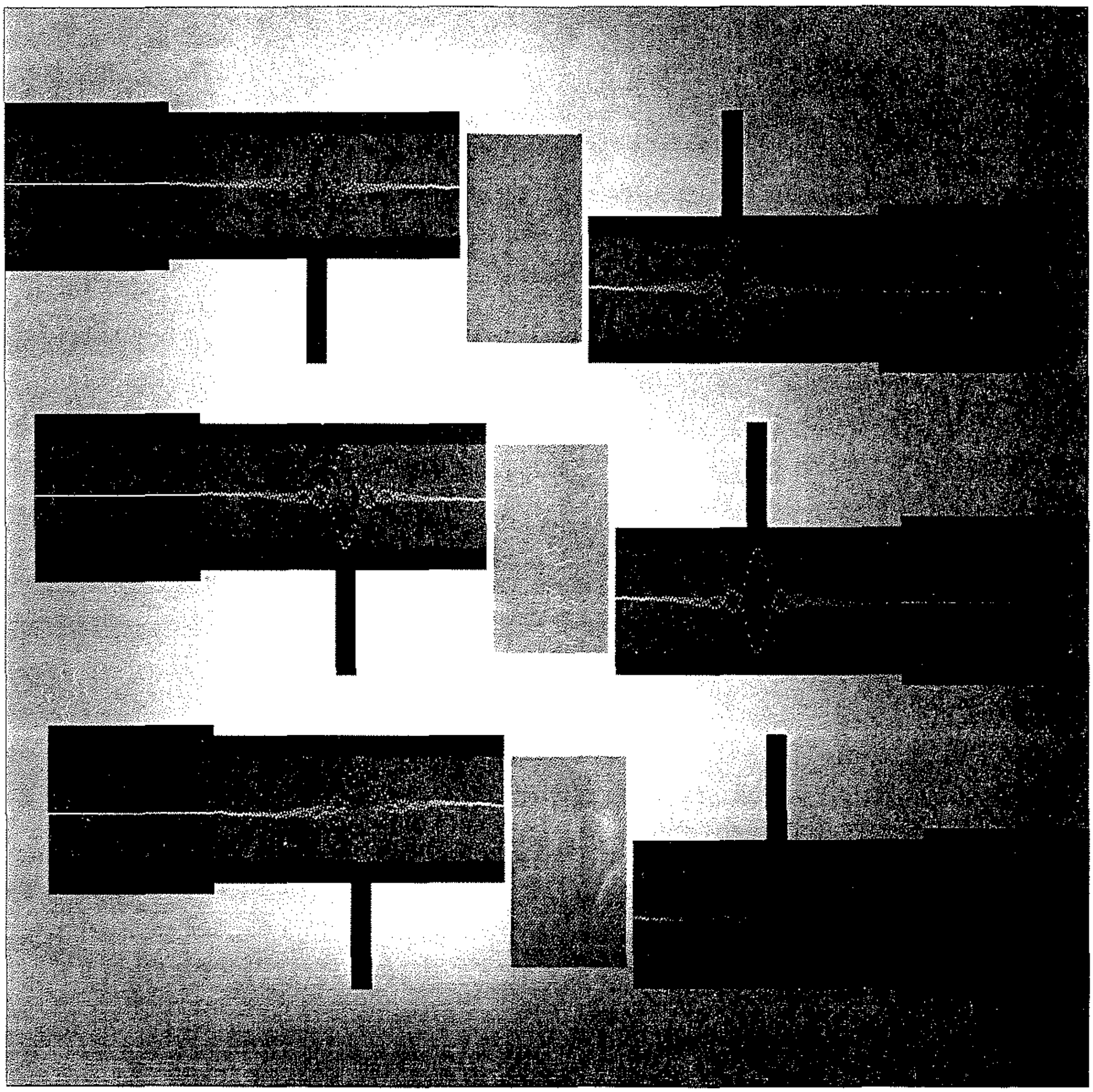
Spectrum Analyzer Applications in Baseband Measurements

New Products

Volume 10
Number 3

1978

Tekscope



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Tekscope

Customer information from
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Beaverton, Oregon 97077

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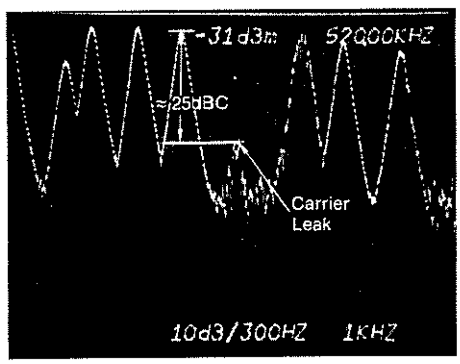
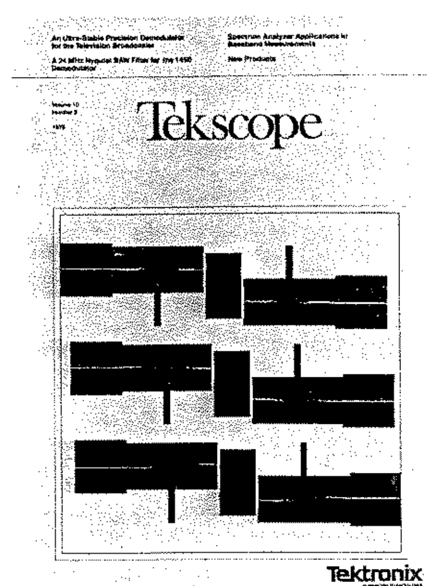


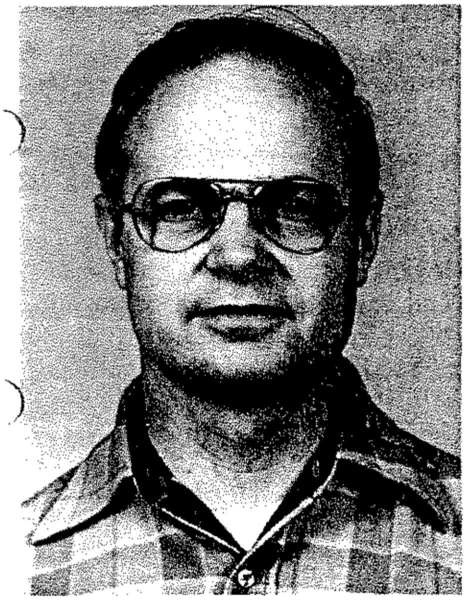
Tekscope is a bimonthly publication of Tektronix, Inc. In it you will find articles covering the entire scope of Tektronix' products. Technical articles discuss what's new in circuit and component design, measurement capability, and measurement technique. A new products section gives a brief description of products recently introduced and provides an opportunity to request further information.

To better serve customers who maintain their TEKTRONIX instruments, the service information formerly appearing in Tekscope will be expanded and published in a publication dedicated to the service function.

Cover: The finger pattern on the SAW filter substrate is fine enough to serve as a diffraction grating as depicted in this photo of a photomask containing three SAW filter patterns.



An Ultra-Stable Precision Demodulator for the Television Broadcaster



Steve Roth has been with Tek since 1965 and has applied his talents to designing television products during this time. He participated in design of the 520 Vectorscope, the 140 Series generators, and the 650 Series color monitors; he is Project Leader for the 1450 Series. Steve has five patents on television circuitry. He received his BSEE in 1964 from Oregon State University.

The quality of the television picture we view in our homes is vastly improved over that of just a few years ago. Technical innovations in both broadcasting and receiving equipment have brought about this improvement.

In the forefront of these developments have been advances in our ability to measure and analyze the television broadcast signal. Precision demodulators play an important role in this measurement capability. As the major link between the transmitted television signal and the baseband (video) measuring equipment, it is essential that the demodulator itself not introduce distortion in the demodulation process. However, three serious problems occur in today's demodulators that make it difficult to achieve and maintain this level of performance—quadrature distortion caused by envelope detection; poor long- and short-term stability of tuned circuits caused by mechanical and thermal shock; and changes in bandwidth characteristics as a function of gain.

In the new TEKTRONIX 1450 Demodulator, these problems have been overcome with new technology and new components. The use of synchronous detectors resolves the quadrature distortion problem. Surface acoustic wave (SAW) filters and low-loss, temperature stable, discrete filter elements minimize the effects of mechanical and thermal shock. And constant gain amplifiers coupled with programmed PIN diode attenuators produce constant bandpass characteristics over a wide range of input signal levels.

Design goals

Versatility and high performance are often considered conflicting goals. Both were set for the 1450. Measurement of signals at the transmitter site and remotely off the air were to be made with no changes occurring in bandpass characteristics. Need for external attenuators was to be eliminated or minimized. Measurement of both visual and aural signals was to be provided for, and video and audio outputs suitable for rebroadcast were to be supplied. Ease of both initial calibration and field maintainability were to present a marked improvement over other available demodulators. These and other goals were coupled with a rigid set of specifications to define a demodulator that would set a new standard for the broadcast industry.

Plug-in down converters

The 1450 consists of two major assemblies—a mainframe containing the power supplies, IF, video, and audio circuitry; and a television down converter (TDC) which is a plug-in front end dedicated to the customer's channel frequency. The Standard Broadcast System M TDC converts the RF input signal to the 45.75 MHz visual IF. An optional System M TDC is also available for the 37 MHz IF, and customizing to other frequencies in that system is possible.

RF input signal levels from −69 dBm to −3 dBm can be fed directly to the input of the TDC. For stronger signals, an attenuator in the mainframe extends the maximum input range (in 10 db steps) to +27 dBm.

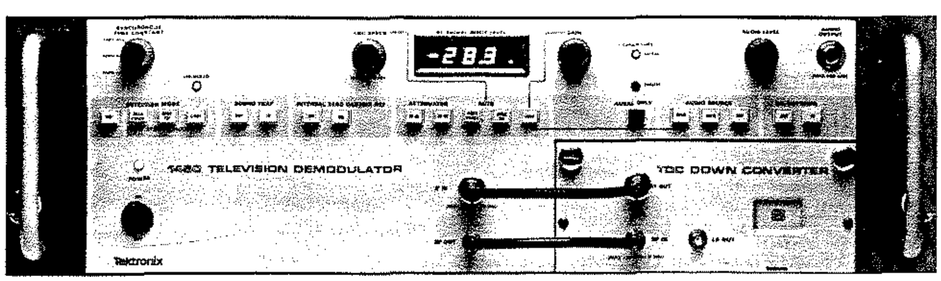


Fig. 1. The 1450 NTSC Television Demodulator. The plug-in down converter customizes the 1450 to any VHF or UHF channel frequency.

The attenuator is a slab-line, thick-film device which, in addition to attenuation, provides a clean 50 Ω load for the incoming signal and a 50 Ω source for the filter that follows.

From the input attenuator, the signal goes to a front-panel connector where it can be easily patched to the RF Input of the TDC. The TDC is housed in a milled aluminum housing with several compartments providing essential shielding between circuits.

A block diagram of the TDC is shown in Figure 2. In the TDC, the signal encounters the first of two ndpass filters tailored to the .annel frequency for which the TDC is designed. Each filter consists of two helical resonators, with adjustable capacitive coupling (see Figure 3). The helical center conductor consists of a selected number of silver-plated tungsten turns which are applied to a ceramic core, using thick-film techniques. The result is a low-loss filter with excellent temperature stability.

The first of three PIN-diode attenuators that make up a calibrated AGC system is located in the TDC. Each attenuator is controlled by a PROM tailored to the particular attenuator and its operating environment. Each provides 21.7 dB of attenuation in 0.7 dB steps. With the 1.4 dB obtained in the analog AGC, a total of 66 dB of attenuation is realized.

Following the PIN attenuator, the signal gets a 16 dB boost from a wideband (50 MHz-900 MHz) amplifier and passes through the second filter to the first mixer. Eight Schottky diodes are used in the ring mixer which converts the RF signal to the IF signal. An 8-diode ring is used instead of the usual 4-diode ring to provide better intermodulation performance.

The first local oscillator (L.O.) is a voltage controlled oscillator (VCO) operating at a frequency equal to the RF input plus the IF. The VCO uses a quarter-wave helical resonator (with the same temperature-stable construction used in the bandpass filters) as the frequency determining element. The use of a VCO and a sampling phase lock loop yields the stability of a crystal oscillator, yet requires only seven crystals to cover all VHF and UHF channels. Transmitters using ±10 kHz offset frequencies are also easily accommodated. The 6-MHz crystals drive a snap-off diode frequency-comb generator which, in turn, drives the phase detector. The helical resonator in the VCO is tuned near the desired L.O. frequency and then locked to the appropriate 6 MHz spur. Pull-in range of the VCO is ≈ 1 MHz with a phase lock loop bandwidth of about 20 kHz.

As a result of the use of a single crystal to lock to more than one channel, the first IF may be off-frequency by as much as 100 kHz for some channels. The mainframe circuitry is designed to handle offsets of up to 250 kHz. (The converter L.O. is adjusted to compensate for such offsets before the SAW filter is encountered.)

The output of the first L.O. is amplified to about +20 dBm before being applied to the ring mixer, to assure good intermodulation performance. A variable attenuator at the output of the mixer provides an adjustment to normalize the TDC's gain for interchangeability.

The IF chain

In the basic 1450 System M configuration, the IF chain is designed to function at the 45.75 MHz visual IF, with 37 MHz being an option. It can also be custom-configured to other IFs for those who wish to measure and analyze the intermediate outputs of an IF-modulated transmitter.

The IF circuitry must handle a wide range of input signal levels (−20 dBm to −64 dBm) and yet maintain a constant bandpass. This is accomplished by operating the amplifiers at a constant gain and providing gain control with variable attenuation between stages. PIN diode attenuators similar to that used in the TDC are located ahead of the second and third IF amplifiers and provide up to 44 dB of AGC. (The attenuator in the TDC adds another 22 dB of attenuation for a total AGC range of 0 to 66 dB.)

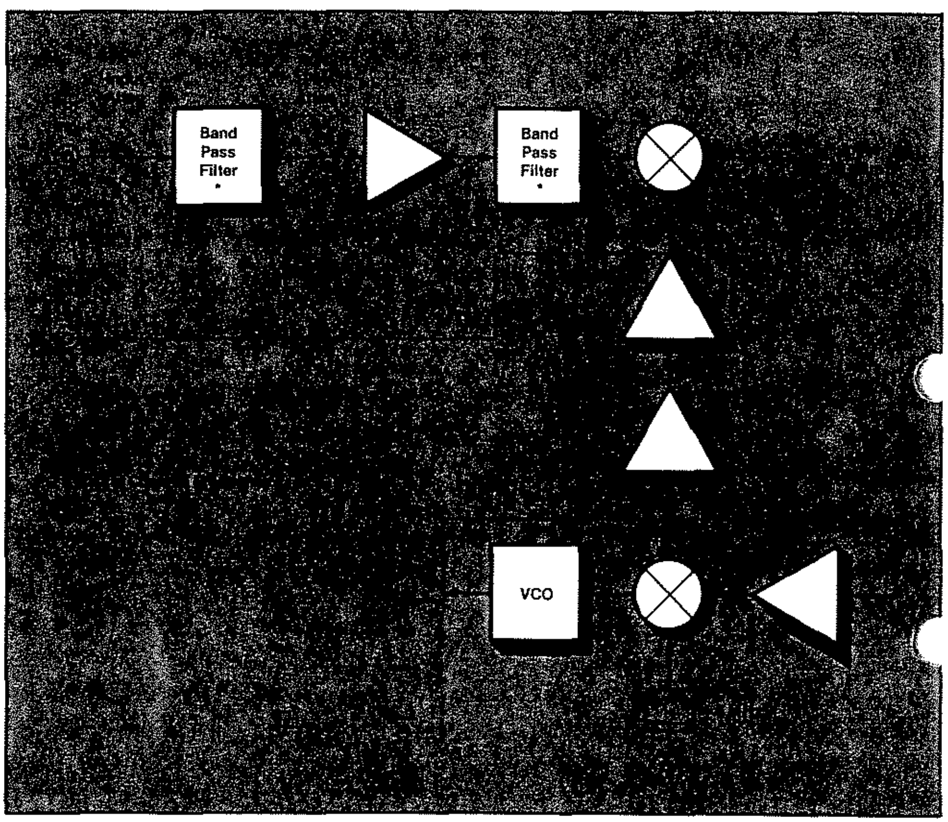


Fig. 2. Block diagram of the down converter. The use of a VCO permits all 83 channels to be converted using only seven crystals. The 0-30 dB input attenuators and AGC attenuators permit operating amplifiers at constant gain over a wide dynamic range of input signals.

The 1450 provides a calibrated digital readout of the input power level, permitting the instrument to serve as a field strength meter with an accuracy of ±1 dB and a resolution of ±0.1 dB. Accordingly, the currents in the attenuator diodes must be set precisely, and differences in diode characteristics must be compensated for. Adjustments are accomplished through digital control of the diode currents. During the calibration process, each attenuator is characterized and the respective values are digitized and burned into PROMs. The PROMs then control digital to analog converters that generate the required diode currents.

A simplified block diagram of the AGC circuitry is shown in Figure 5. Selection of AGC, and selection of back porch or sync tip as the AGC reference level are accomplished by means of front panel controls. For AGC, the video output level is sampled at the selected time and applied to a tracking analog-to-digital converter. The output of the converter drives the decoder which, in turn, controls the PIN diode drivers. Both the PROMs and the digital-to-analog converters are contained in the PIN diode drivers. Fine AGC is applied to the IF post amplifier to fill in between the 0.7 dB steps of the PIN diode attenuators.

The speed of the AGC loop can be set also by a front-panel control to allow the operator to either observe (SLOW mode), or eliminate (FAST mode), variations in input signal levels such as hum modulation or airplane flutter.

AGC circuitry also supplies control signals to actuate alarms in case of loss of the visual or aural carrier.

After the gain control section, the IF signal is converted to 24 MHz and filtered to remove extraneous mixer output signals. The IF converter mixer is a conventional ring diode mixer, with care taken to maintain a high degree of balance in the mixer.

The filtered 24 MHz signal goes to the SAW filter preamplifier, and to a pick-off amplifier having relatively high input impedance, to provide the Aural IF signal.

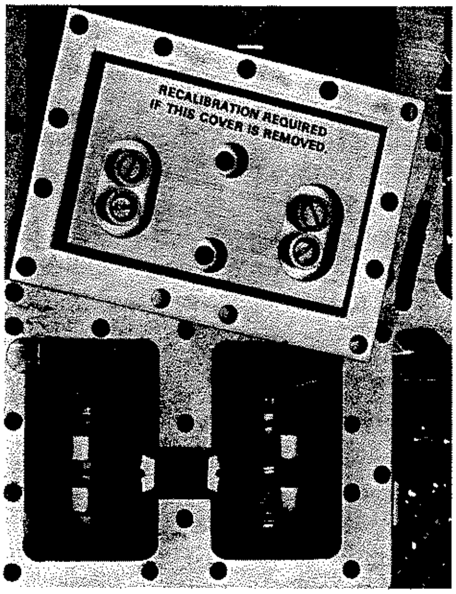


Fig. 3. The RF input bandpass filter consists of two helical resonators capacitively coupled. The helical center conductor is a series of silver-plated tungsten turns applied to a ceramic core using thick-film techniques. The result is a low-loss filter with excellent temperature stability.

The SAW filter

Some of the most significant improvements in performance achieved by the 1450 result from the use of SAW[1] filters to obtain the desired Nyquist slope characteristics. Conventional discrete-element Nyquist filters are complex devices difficult to adjust and maintain.

Two Tek-designed and manufactured SAW filters are used in the IF chain. A wideband unit is used for making out-of-service measurements with only the visual carrier on. A narrow band unit (which attenuates the aural carrier by greater than 50 dB) is used for making in-service measurements with the aural carrier on.

SAW filters offer several advantages over discrete-component designs: The desired bandpass characteristics are more easily achieved. For example, in the 1450, the Nyquist slope incorporates the S-shape characteristics which are in the new Demodulator Standards being prepared by the EIA. (see Figure 6). The filter requires much less space. There are no calibration adjustments. And the filters' characteristics do not change with mechanical or thermal shock.

However, along with the advantages, the SAW filter concept offers several design challenges. While relative response characteristics remain constant with changes in temperature, the absolute frequency does not. A SAW filter is also difficult to drive. It requires low driving and load impedances (≈10 Ω) and has an insertion loss of about 30 dB. This insertion loss, and the fact that the ultimate rejection desired is greater than 60 dB, means that greater than 90 dB of isolation between input and output is required to achieve the desired 1450 Demodulator performance. (This isolation is equivalent to putting in a megawatt of power and being able to receive only one milliwatt.)

The excellent performance of the SAW filter made the design challenges worth tackling. The temperature-induced change in operating point was resolved by placing a temperature sensor near the filter to generate a correcting signal for the converter L.O. And extensive shielding techniques at the input and output of the filter eliminated the undesired coupling.

The phase lock section

A simplified block diagram of the phase lock section is shown in Figure 7.

The VCOs for the converter L.O. and the reference L.O. have to meet two conflicting requirements—low phase noise generation and wide pull-in range capability. Phase noise must be low because any phase variations in these oscillators are added directly to the overall detected signals. This could obscure the phase measurements that the 1450 can provide. Wide pull-in range is needed to accommodate the ±100 kHz variation in the incoming IF, and also to thermally track the SAW filter.

The response time of the reference L.O. phase-control loop is made

[1] See article entitled "A 24 MHz Nyquist SAW Filter for the 1450 Demodulator" in this issue.

selectable so that phase errors in the incoming signal can be displayed and measured (SLOW mode) or tracked out and either eliminated or reduced (FAST mode).

The VCOs that comprise the converter L.O. and the reference L.O. are of similar design. Both are composite oscillators, combining the low phase noise of a crystal-controlled oscillator (XCO) with the wide pull-in range of an LC oscillator.

In both cases, the VCO output is compared to that of an XCO, and the resulting difference is converted to a correction signal by a pulse-count discriminator. The discriminator aids in achieving a rapid lock by supplying a large correction signal when the VCO is considerably off-frequency, and a progressively smaller correction signal as lock is achieved. This correction signal is used to balance out the voltage control signal.

The frequency lock circuitry of the reference VCO also accepts a correction signal from a temperature sensor to allow for temperature tracking of the SAW filter characteristics.

The reference L.O., through the temperature compensation, is kept at exactly the frequency to which the incoming IF signal must be converted, to pass through the SAW filter properly. The limiter output, because it is the same frequency as the converted IF signal, is then compared to the reference L.O. Any frequency difference between the two signals is representative of the frequency shift that must be obtained from the converter L.O. to bring the converted IF signal "on frequency".

The frequency lock system, however, does not have the capability of responding to fast phase disturbances in the incoming RF signal. The reference L.O./converter L.O. phase lock must work through the SAW filter which has about 7 microseconds of delay. This limits the rate at which corrections can be applied to that loop.

This difficulty is overcome by providing a method of shifting the phase of the reference L.O. so that it can track phase irregularities in the visual signal. Absence of delay or storage elements in this control loop allow the phase to be changed as rapidly as desired.

The correction signal for the phase shifter is derived by sampling the output of the quadrature detector during some "resting time", such as backporch or sync tip (front panel selectable). Since the output of the quadrature detector should be zero at those times, it can be used as the control signal for the phase shifter. (A continuous mode of correction is also available by front panel selection, if desired.) The control loop will adjust the phase of the reference L.O. to make the output voltage of the quadrature detector be zero at the selected time.

The limiter

The limiter plays an important role in the second L.O., video detector, and aural functions and has some stringent requirements. It must accommodate a wide range of signal levels (up to 40 dB with modulation), yet introduce less than one degree of phase shift, at frequencies approximating 24 MHz.

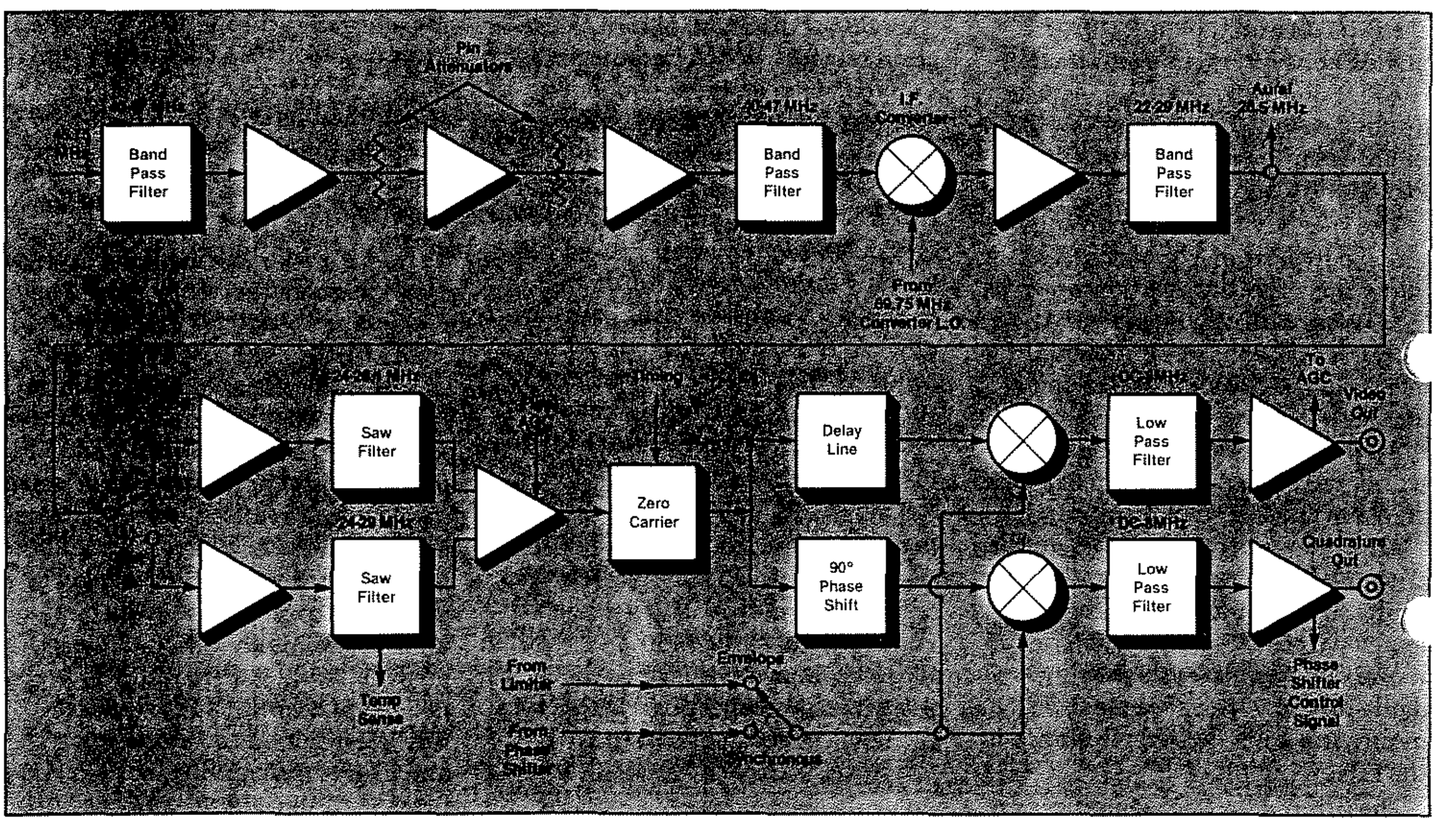


Fig. 4. Block diagram of IF and synchronous detector. Multiple amplifier stages are used to accommodate 0 to 66 dB range of AGC attenuation. Separate SAW filters are used for "visual only" or "visual plus aural" operations.

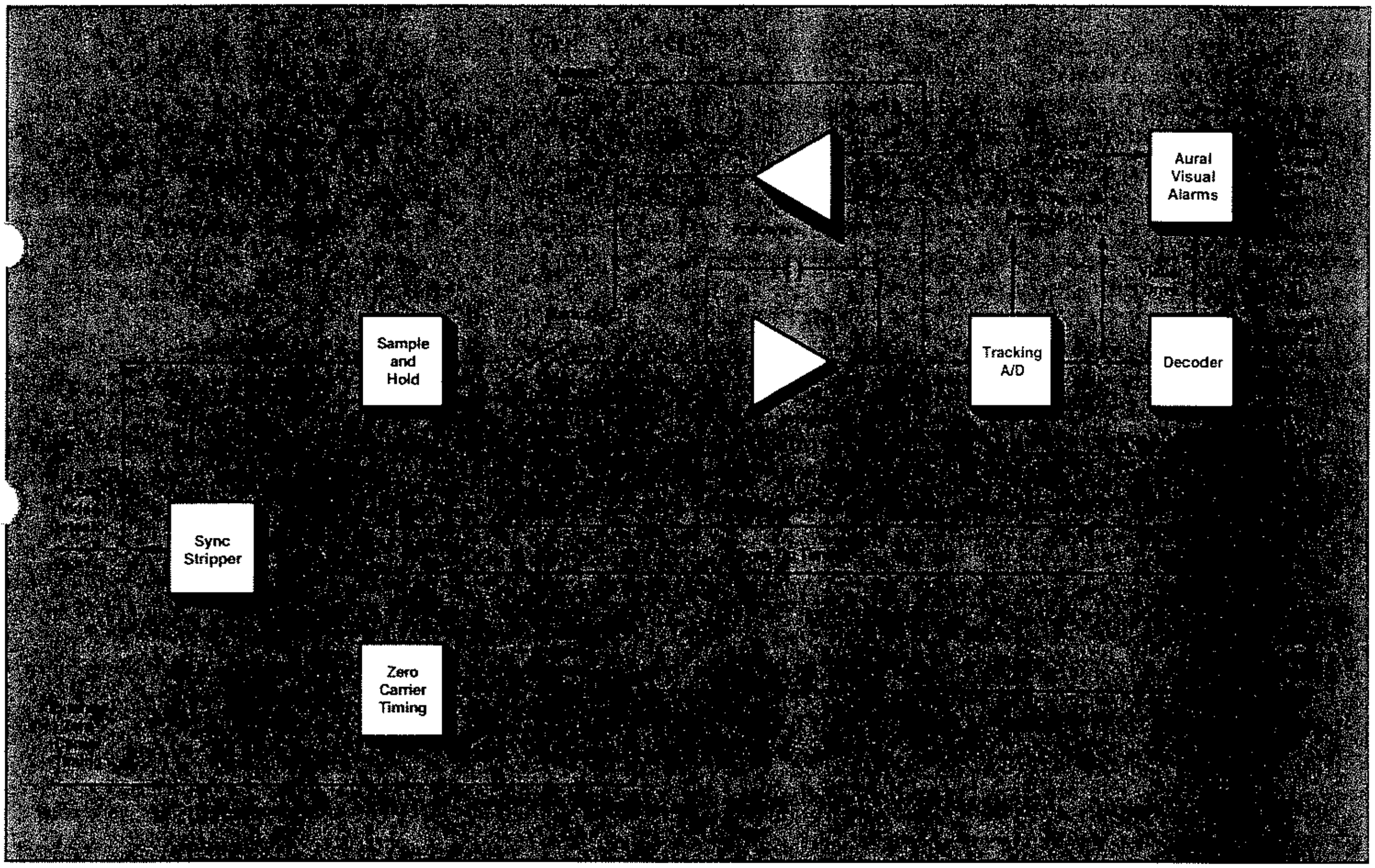


Fig. 5. AGC circuitry drives three PIN-diode attenuators for AGC range of 0 to 66 dB. PROM control of attenuator current yields readout accuracy of RF input within ±1 dB.

Four differential amplifier stages provide a total gain of 60 dB. Adjustable current sources for the amplifiers provide a delay adjustment mechanism for the limiter.

With amplitude variations removed, the limiter output can also serve as the local oscillator for the envelope detector. And since any transmitter phase noise is inherently present in the limiter output, the limiter output is used as the local oscillator for the first aural mixer when analyzing intercarrier IF performance.

The synchronous detector

The synchronous detector consists of two product detectors—one supplying the video output, the other the quadrature output. The IF signal to the video detector passes through a delay line of approximately 10.4 ns, which corresponds to 90 degrees of the 24 MHz IF frequency. The IF signal to the quadrature detector, on the other hand, passes through a bandpass filter

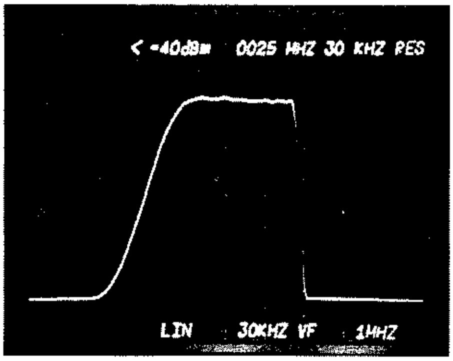


Fig. 6. A "textbook" Nyquist slope response is achieved by the SAW filters used in the 1450.

which has the same 10.4 ns of delay at 24 MHz, but introduces no phase shift of the IF carrier. Thus, two signals in time coherence but phase quadrature are produced to be detected by the product detectors.

The detector/amplifier combinations have a stringent stability requirement because the quadrature output is the control signal for the phase lock. Any errors in that signal, such as those caused by thermal drift, will cause the phase coherence between the IF signal and the reference L.O. to be in error. Since accurate phase measurements must be made and good transient response maintained, the phase error cannot be allowed to exceed 1 degree. This means the quiescent output of the quadrature detector/amplifier combination must not change by more than ±10 mV over the operating temperature range. This kind of performance is typically achieved only with chopper stabilization techniques.

In the 1450, the required stability was achieved with innovative circuit design. An integrated-circuit doubly-balanced mixer is used as a current mode switching detector. The bias current of the IC is set up to change with temperature, to maintain constant transductance. The transductance is stabilized if the dynamic emitter resistances ($r_e = \frac{mk\,T/\alpha}{I_E}$) are held constant, which is done by allowing I_E to cancel out the effects of the T (temperature) term.

The output amplifier that follows the detector has been subjected to equally concentrated design. Since it is a feed-back amplifier, it is inherently more stable than the detector, which must exhibit "open loop" stability. The input offset voltage, however, is thermally tracked to compensate for amplifier output voltage changes caused by temperature variations.

Further, to insure that the potential stability is actually achieved, the instrument is subjected to environmental temperature cycling in the calibration process, during which a single compensating resistor is selected and installed.

The same detector/amplifier circuitry used in the quadrature channel is also used in the video channel. As noted earlier, the video output signal is sampled and used as the control signal for the AGC system. Any errors in its output will affect the gain stability of the instrument.

The end result of this attention to design in the detector/amplifier stages is a quadrature output well within the 1-degree specification and a video output level held within 1 percent by the AGC.

The correction voltage from the quadrature output is shaped by diode matrixes and applied to two mixers driven 90 degrees apart by the 24 MHz reference oscillator. The dc levels to the mixers are such that with zero output from the quadrature detector, the output of one mixer starts at full amplitude. As the correction voltage to the mixer increases, the output falls off in a sinusoidal manner until it reaches zero. As the correction voltage is decreased, the output falls off in the opposite direction until it reaches zero. These are the two limits of the control signal.

The correction voltage to the other mixer is such that its output starts out at zero amplitude and increases in amplitude as the correction voltage is increased, and vice versa.

The outputs of the two mixers are combined to give a constant amplitude sine wave whose phase can be shifted linearly with voltage. The bandwidth of this system is such that corrections can be made at a line rate, which is actually a limit imposed by the sampling of the reference time (sync tip or back porch).

The audio system

With an audio section bandwidth of 30 Hz to 20 kHz, one of the potential applications for the 1450 is as an off-the-air receiver for rebroadcast purposes.

The 28.5 MHz aural IF carrier is converted to 4.5 MHz using either the 24-MHz phase lock reference oscillator ("Split" mode), or the limiter output ("Intercarrier" mode), as the first L.O. Amplified and filtered, the 4.5 MHz signal is then converted to 1 MHz using a 5.5 MHz crystal oscillator for the second L.O. A limiter removes any amplitude varia-

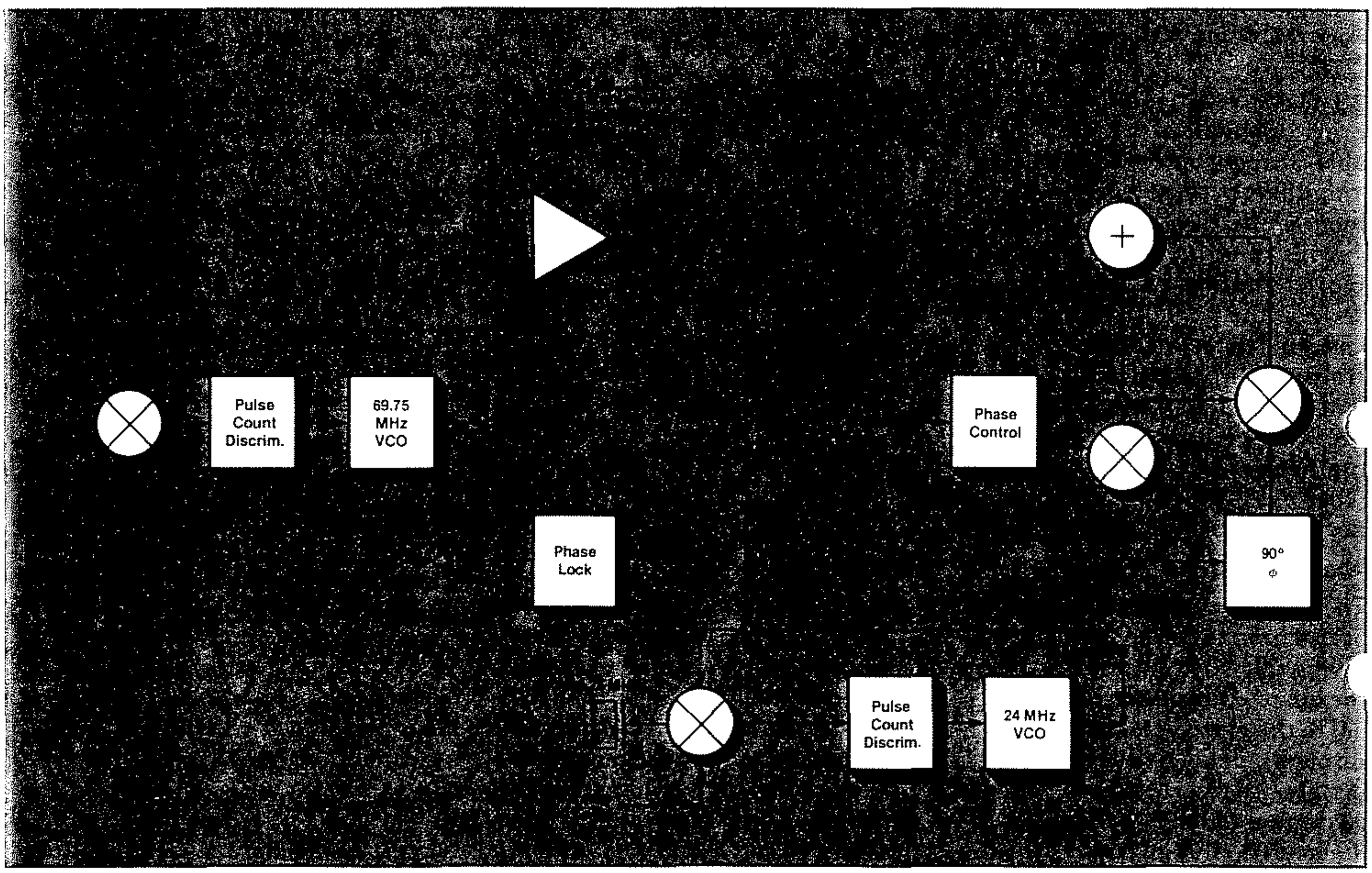


Fig. 7. Converter L.O. and phase lock circuitry. Two mode system uses a frequency control loop to give a wide pull-in range, and a phase control loop for fast phase correction.

tions, and the 1 MHz signal is then demodulated using a pulse count discriminator.

The pulse count discriminator is operated at 1 MHz to increase the available output signal, thereby improving the signal-to-noise ratio over the same circuit operating at 4.5 MHz. The discriminator puts out pulses of constant amplitude and duration, whose repetition rate varies at the modulating frequency. The discriminator is an FM detector which exhibits high linearity, and is the main factor in achieving the low 0.2% maximum harmonic distortion specification.

Multiple audio outputs are provided, including 600-ohm balanced line, 8-ohm speaker, and front-panel headphone jack. Bandwidth of the audio section is 30 Hz to 20 kHz. Other outputs include an aural alarm to indicate loss of the aural carrier, a calibrated deviation output, and a 4.5 MHz IF output.

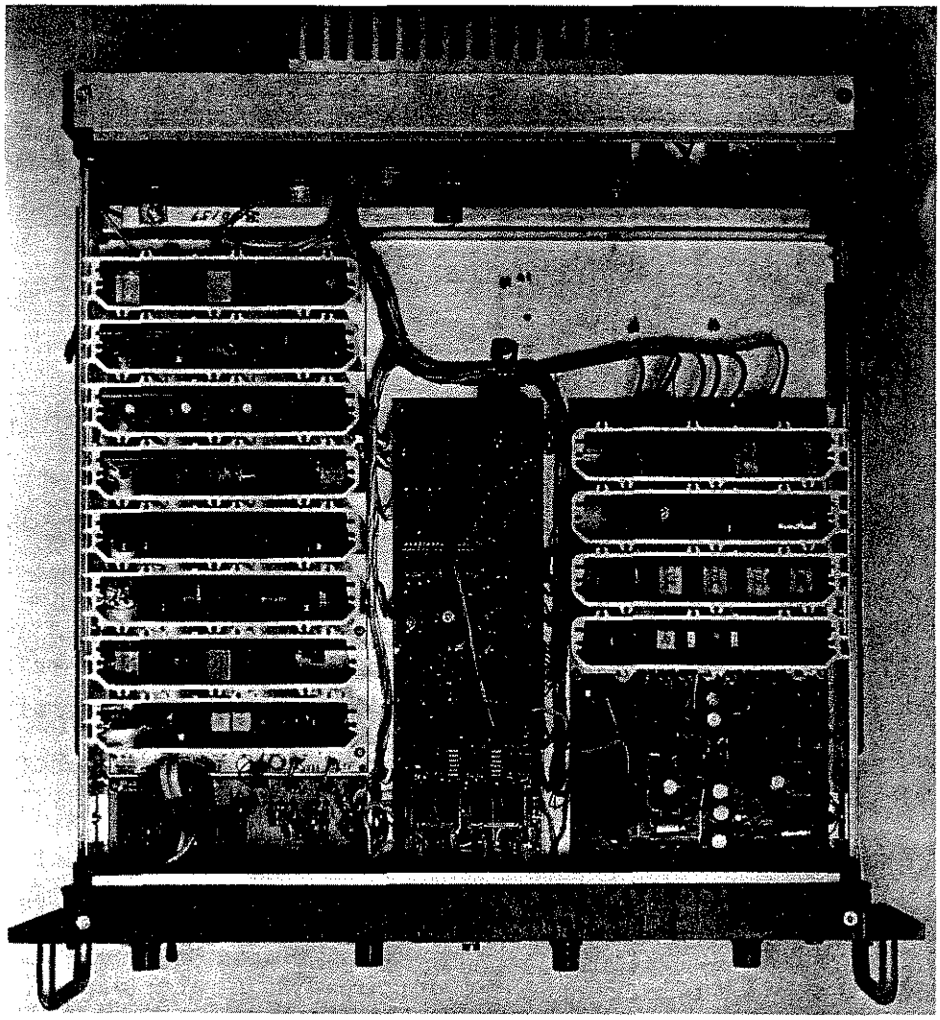


Fig. 8. The major part of the circuitry is mounted on small circuit boards housed in extruded aluminum compartments. Separate covers for each compartment enhance isolation.

Mechanical design

The 1450 interior is shown in Figure 8. Small circuit boards housed in extruded aluminum compartments comprise the major circuitry. The compartments provide essential shielding between the many oscillators and the sensitive circuitry. Individual covers for the compartments give maximum isolation. Interconnection between circuit boards is accomplished with interface boards and a few rigid coaxial lines. Easy access to components is provided by use of a circuit board extender.

Acknowledgements

Many people, in many departments, contributed to the successful completion of the 1450 project. Project Manager Chuck Barrows provided overall direction, with Charlie Rhodes challenging us with the seemingly impossible while providing many valuable inputs. Assisting in electrical design were Jim Zook, front-end design (including the ceramic core resonators); Keith Ericson, IF and synchronous detector; and Dan Nicholas, AGC and audio. Bill Drummond, Tek Labs, was responsible for the SAW filter designs; and Oscar Olson, Hybrid Circuits Engineering, worked with Jim Zook on the helical resonators. The front-end precision attenuators are the work of Jack Roberts, Frequency Domain Instrumentation. Paul Chadwick handled the mechanical design. Much credit is also due Ellis Workman for his circuit board layouts, and Bob Thacker in smoothing the transition to manufacturing. Prototype support technicians were Sherry Self and Kerry Montgomery.

A 24 MHz Nyquist SAW Filter for the 1450 Demodulator

Bill Drummond came to Tek from the S.F. Bay Area in '73, starting in the [illegible] group. In late '74, he joined the Instrument Research Group working on SAW filter design. Bill played a major role in development of the SAW filters used in the 1450. He received a BSEE from Healds Engineering College in [illegible].

One of the major problems experienced by the users of precision broadcast demodulators is the change in the Nyquist filter's response caused by temperature change and mechanical shock. To achieve a reasonable Nyquist response requires a filter with many elements. Such filters are time-consuming to adjust and difficult to keep in adjustment. Yet, a proper Nyquist response is essential to the demodulator's function.

This presents an ideal application for a surface-acoustic-wave (SAW) device. These devices are usually small in size, highly reliable, and their phase response can be specified independently of their magnitude response, a distinct advantage in this application. They can produce a frequency response that is difficult, if not impractical, to achieve using discrete elements; and once designed, the devices are easily reproducible.

Perhaps a review of SAW filter basics would be useful at this point. Surface acoustic waves are essentially what the name implies. They are waves that travel along the surface of a polished substrate, penetrating only a very small distance into the surface of the material. Scientifically they are called Rayleigh waves, in honor of Lord Rayleigh who first described the acoustical waves that travel along the earth's surface after an earthquake.

Nearly any material with low mechanical losses such as glass, quartz, ceramics, and some types of metal will support the propagation of surface waves. The mechanism most often used to generate the surface wave is the piezoelectric effect. Accordingly, single crystal piezoelectric materials such as quartz and lithium niobate are generally used as the substrate for SAW devices.

The substrate is typically about two millimeters thick, with the length determined by the desired frequency response. Narrow filter bandwidths at low frequencies, or very steep skirts at any frequency, require long transducers and, thus, long substrates. Surface wave filters are currently in use from 10 MHz to 1.5 GHz. The lower limit is set by the practical size limits on the substrates, while the upper limit is set by the resolution of state-of-the-art lithography techniques and substrate velocities. One semiconductor manufacturer recently announced production of 3 GHz SAW units using two-stage electron-beam lithography.

The drawing in Figure 1 illustrates generally how the surface wave affects the substrate material and how its effect decreases as a function of depth into the material. Both longitudinal and transverse components are present in the surface acoustic waves. The velocity of the acoustic wave is typically five orders of magnitude slower than the velocity of light, which accounts for the frequent use of SAW devices as delay lines. For example, an electrical signal takes about three microseconds to traverse a cable one kilometer long. The acoustic wave travels about one centimeter in the same length of time. Hence, equivalent delays can be achieved with a substantial reduction in space and weight.

Excitation and detection of acoustic waves is accomplished by depositing an input and output transducer on the piezoelectric substrate using thin-film techniques. The transducers consist of two electrodes in the form of interdigital fingers. (See Figure 2). The spacing between the fingers determines the frequency of the acoustic wave that will be generated.

Applying a periodic voltage of the appropriate frequency to the input transducer causes a corresponding displacement to occur in the piezoelectric material. This periodic displacement propagates away from the input transducer in the form of surface waves in both directions normal to the fingers of the transducer. As the surface waves travel

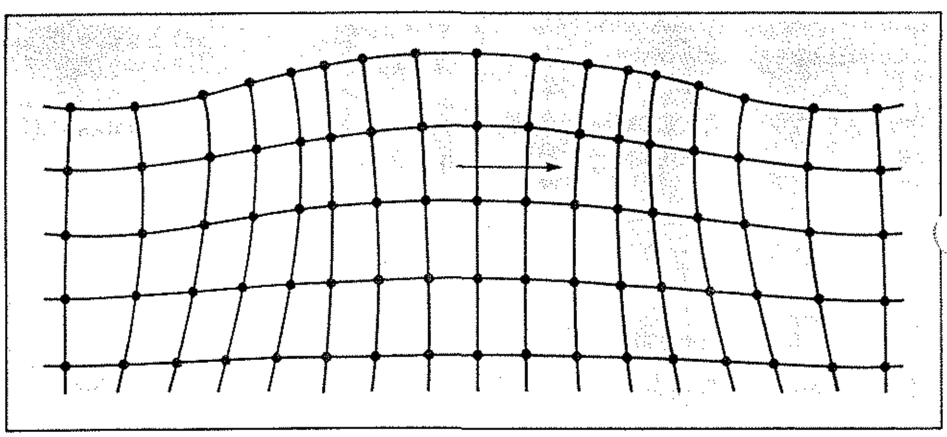


Fig. 1. A cross-sectional view of the substrate illustrating how the substrate material is affected by passage of a surface acoustic wave.

down the substrate, they encounter the output transducer. If the periodicity of the surface wave and the fingers of the output transducer match, an output voltage is generated across the output transducer.

Any filter's frequency response is uniquely related to its impulse response through the Fourier transform. SAW filters are realized by arranging the filter transducers so that they produce the impulse response related to the desired frequency response. Transducers can be made to produce almost arbitrary, finite, impulse responses by varying finger lengths as a function of position along the transducer. For example, to produce a rectangular bandpass characteristic, its Fourier transform (*sin*x/x) is designed into the transducer pattern.

One of the important properties of SAW filters is the simple metal-on-insulator, one-mask fabrication process. Once the mask is defined, device reproducibility is a function of the substrate characteristics, metal quality, and metal thickness—all of which can be made highly consistent. This means devices having a precise frequency response are easily reproducible—a substantial improvement over complex discrete element filters with their inherent adjustment and stability problems.

There are, however, some tradeoffs involved in the use of SAW filters—one of the most serious being insertion loss. It was noted earlier that conventional biphase transducers radiate half the electrical-to-acoustical conversion in each direction; conversely, the output transducer can, at most, reconvert into an electrical signal only half the acoustic power incident on it, for a minimum insertion loss of 6dB.

Another consideration is the effect of triple-transit echoes caused by a part of the energy being reflected by the output transducer and re-reflected by the input transducer. This produces a periodic ripple in phase and amplitude in the passband of the filter. The triple-transit signal can be substantially reduced by electrical mismatching of the input and output transducers. However, this also reduces the level of the desired signal, which then must be amplified, with attendant deterioration in the signal-to-noise ratio.

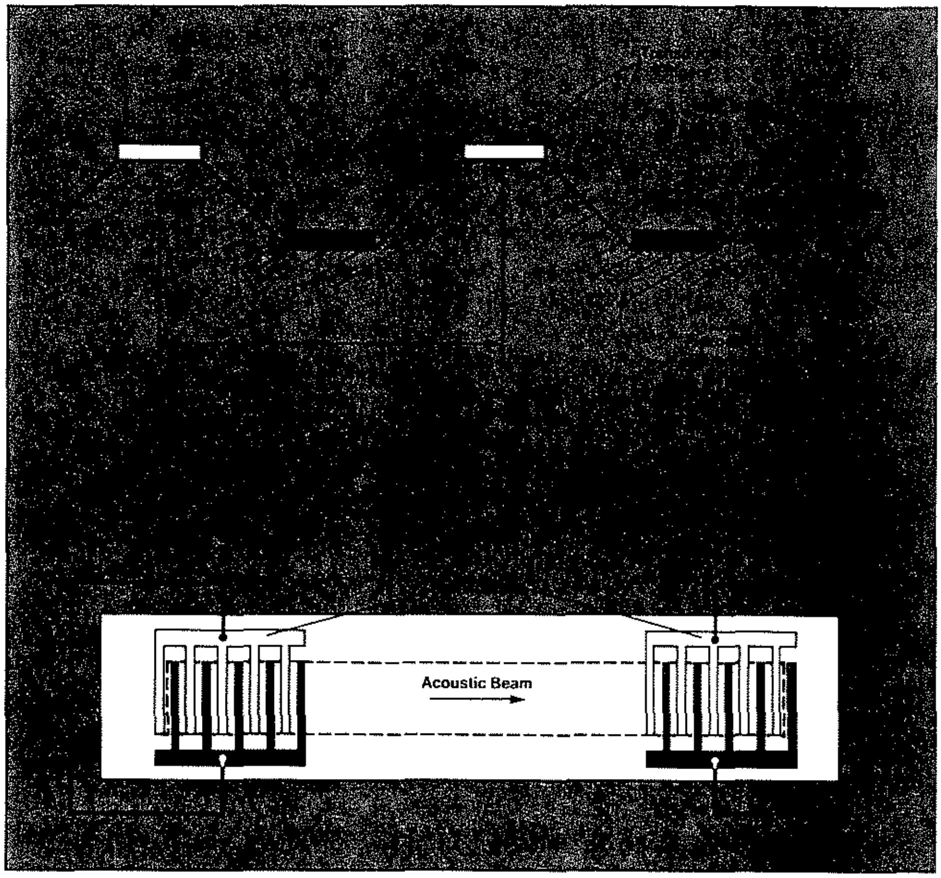


Fig. 2. An illustration of surface acoustic wave transducer operation. The spacing between pairs of fingers determines the frequency of the acoustic wave that will be generated. The input and output transducers are identical in design.

A final consideration is the bulk wave effect caused by acoustic energy transmitted through the bulk of the substrate and reflected from the edge surfaces. This can be minimized by the use of damping material to mount the substrate, the use of a track changer, and other techniques, but it is difficult to eliminate completely.

Some Uses for SAW Devices

Before leaving the discussion of SAW filter basics we should mention some of the uses for SAW devices. Bandpass filters and delay lines are the two most common applications. Delay times from a few tenths to several tens of microseconds are eas-

(A)

(B)

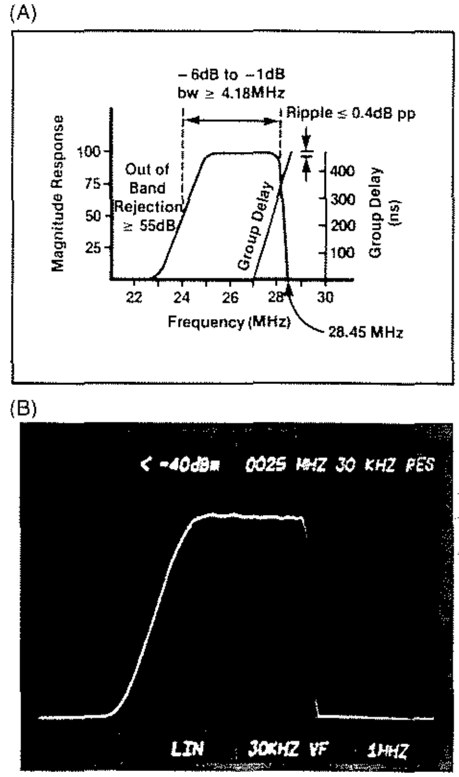


Fig. 3. (A) Desired, and (B) actual response of the 1450 narrow band filter. Both curves are in terms of a linear vertical scale.

ily obtainable. Delays up to 20 milliseconds have been achieved by passing a signal in a closed loop or helical path around the crystal.[1] However, these millisecond devices are still in the laboratory stage.

The most important uses of delay lines to date have been for demodulation functions for receivers—particularly, differential phase shift keying demodulation, frequency discrimination for FM receivers, and noncoherent moving-target-indicator radar receivers.

Recently-developed SAW resonators are finding increasing use as the frequency determining element in VHF and UHF oscillators. Their small size and high Q simplify amplifier design and provide better stability and immunity to external component variations.[2]

Other uses for SAW devices are still in the development stage—frequency synthesizers, matched filters for correlation applications, and electronically variable delay to name just a few.

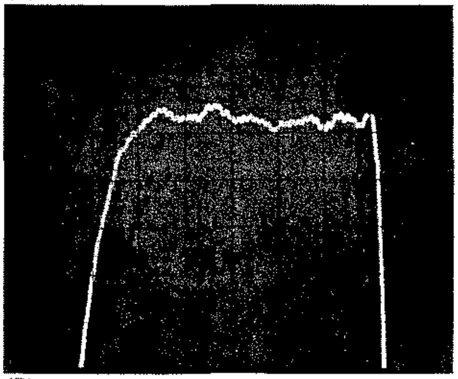
The 1450 SAW Filters

With this brief background, let's consider the SAW filters designed for the 1450. There are two of them—a wide-band unit with a passband of 24-29.5 MHz, and a narrow-band unit (24-28.1 MHz) which includes a sound trap to remove the aural carrier.

The narrow band filter has some unique characteristics as shown in the illustration of the desired response in Figure 3(A). The actual response achieved is shown in the photo in Figure 3(B). Both curves are in terms of a linear vertical scale. The actual response is also shown in Figure 4 with the vertical scale in terms of dB. The top photo shows the peak-to-peak ripple to be well within the 0.4 dB specification, while the bottom photo shows the filter out-of-band characteristics to be significantly better than the −55dB specification.

[1]Scientific American, October 1972, "Acoustic Surface Waves."
[2]Proceedings of the IEEE, Vol. 64, No. 5, May 1976. "Surface Acoustic Wave Devices for Communication."

(A)



(B)

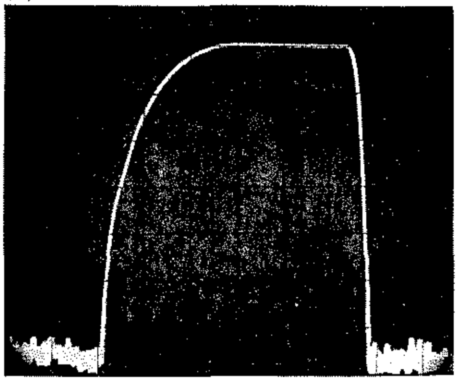


Fig. 4. The actual response with the vertical scale in terms of dB. (A) Top photo shows ripple to be well within 0.4 dB specification. Vertical sensitivity = 0.25 dB/div and span is 500 kHz/div. (B) Out-of-band characteristics are better than −55 dB specification. Vertical sensitivity = 10 dB/div and span = 1 MHz/div.

Both filters use a lithium niobate substrate. This material has a higher temperature coefficient than quartz but is much more efficient in coupling energy between the transducer and the substrate. With a change in temperature, the shape of the response curve remains the same but the center frequency shifts. This shift is compensated for in the 1450 by placing a temperature sensor near the SAW filter to generate a correcting signal for the IF converter oscillator.

By SAW filter standards, the physical size of the devices used in the 1450 is large. Packaged size is 1.4 x 2.3 inches. This is due to the relatively low frequency (24 MHz) and the steep skirt required for the sound trap. The cover photo shows the general layout of the SAW filter transducer patterns. The series of parallel unconnected fingers in the center of the device constitute a track changer called a multi-strip coupler. The track changer translates the acoustic energy from the input transducer acoustic track to that of the output transducer. Since the track changer is relatively insensitive to the bulk wave energy generated by the input transducer, little bulk wave signal is seen by the output transducer. To reduce bulk wave effects still further, the substrate is mounted in energy-absorbing material. The completed filter is then housed in a hermetically sealed package back filled with dry nitrogen to keep out dust and moisture.

In the circuitry surrounding the filter, both the input source and output load termination impedances are deliberately mismatched to the electrical impedances of the input and output filter transducers. Although the mismatch increases the overall insertion loss, it is required to minimize triple-transit-reflection effects. The losses involved in the track changer, losses due to bidirectionality, and losses due to mismatching add up to an insertion loss of about 30 dB for the SAW filters used in the 1450. This places some stringent requirements on the output amplifier of the SAW filter to maintain a favorable signal-to-noise ratio. The advantages realized by use of the SAW filter, however, far outweigh these considerations.

The SAW filters complement the precision and stability designed into the remainder of the 1450 circuitry, to create a demodulator which sets a new standard for the broadcast industry.

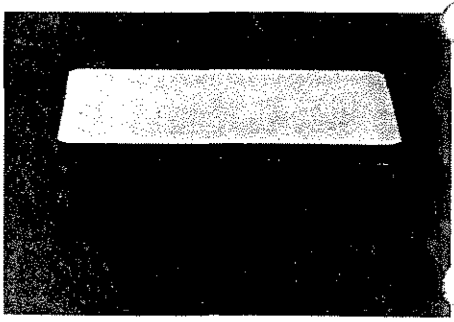
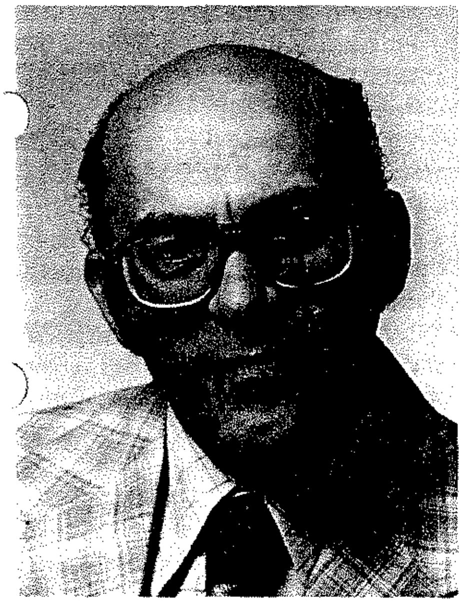


Fig. 5. Hermetically-sealed SAW filter package used in the 1450. Package is 3.6 cm by 5.8 cm.

Spectrum Analyzer Applications in Baseband Measurements



Morris Engelson has authored numerous articles on the design of spectrum analyzers and their application. He is well qualified to write on this subject having worked with spectrum analyzers for over 15 years. Morris received his BSEE in 1957 and MSEE in 1962, both from C.C.N.Y.

The modern spectrum analyzer has long been used for frequency division multiplex (FDM) baseband measurements such as hunting for spurious responses or checking intermodulation. This instrument can also be used for most of the measurements usually performed with a selective level meter.

The measurements that a modern spectrum analyzer such as the TEKTRONIX 7L5 can make include:

a. Amplitude levels of carriers, test tones, signal tones, data, etc.
b. Spurious signal levels and frequencies due to harmonics and intermodulation.
c. Leakage at channel carrier, group carrier, or other frequencies.
d. System noise levels.
e. Frequency shifts due to changes in degree of modulation or other causes.
f. Identification of random transient noise burst interference.
g. Determination of notch filter shape.
h. Checking for harmonic distortion.
i. Checking subsystem transmission characteristics by means of a tracking generator.
j. Noise measurements, including C-message and psophometric weighting.

The purpose of this article is to provide the reader with short illustrations of some of the above measurements. Detailed measurement techniques, interpretation of results, and a description of the instrumentation used are available in Tektronix applications literature referenced at the close of this article.

Wide span search measurements

Suppose one wishes to check or adjust the relative levels of the twelve carriers spaced at 4 kHz intervals bounded by 659 kHz and 615 kHz.

There might be other signals within the selected range, such as a pilot tone usually at 656.08 kHz. The simulated channel bank shown in Figure 1 displays only the unmodulated carriers.

With the measuring instrument operating normally with digital storage display, we can observe the twelve channels of interest. Note that the maximum amplitude variation is 2 dB.

Figure 2 shows a search of the full 5 MHz spectrum. A number of active channels below about 2 MHz are shown blending into each other. Two tones occur just above 2.5 MHz. The remainder of the spectrum display up to 5 MHz shows only system noise.

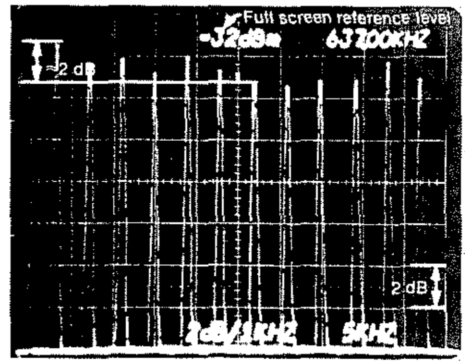


Fig. 1. Spectrum display of unmodulated carriers for 12 channel bank.

Figure 3 shows all baseband activity for the system of interest. One can observe at a glance which groups are occupied, which are missing, and whether anything is there that doesn't seem to belong, such as a suspicious tone at 1606.5 kHz, subsequently identified as spurious.

Harmonic distortion measurements

The 80 dB on-screen dynamic range of the TEKTRONIX 7L5 enables one to measure harmonic distortion down to 0.01%. When a single harmonic is involved, the measurement simply consists of determining the dB level with respect to the fundamental. This is also known as "dB down".

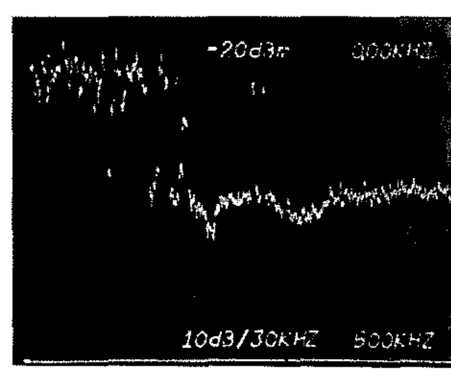


Fig. 2. Zero to 5 MHz display of baseband spectrum.

The dB down level is then converted to a voltage ratio which, when multiplied by 100, yields percent distortion. A simple rule to remember is that distortion changes by a factor of ten times for every 20 dB change in level. Thus, at 10 dB/div and the fundamental at full screen, distortion products 2 divisions down represent 10% distortion, those 4 divisions down represent 1%, and those 6 divisions down represent 0.1%.

Figure 4 shows a −10 dBm fundamental and over ten of its harmonics. This display was taken with a 7L5 Option 25 in a 7603 Mainframe. The instrument parameters are indicated on crt readout. The left upper position (set at 0 dBm) is tracking generator output—not used in this measurement.

We observe the following amplitude levels:

Fundamental	−10 dBm
2nd harmonic	−53 dBm
3rd harmonic	−64 dBm
4th harmonic	−60 dBm
5th harmonic	−71 dBm
6th harmonic	−65 dBm
7th harmonic	−76 dBm
8th harmonic	−68 dBm
9th harmonic	−78 dBm
10th harmonic	−71 dBm
11th harmonic	−85 dBm
12th harmonic	−72 dBm

Intermodulation distortion measurements

The ability to set frequency with a high degree of accuracy makes it easy to intercept intermodulation and other spurious signals whose frequency can be predicted. The high degree of accuracy also means that the frequency of unexpected spurious responses can be pinpointed and the source identified.

If three tones at 57 kHz, 2600 kHz, and 2714 kHz are fed to the system under test, the following second and third order intermodulation responses will occur:

$$a. \begin{cases} 2(57) = 114 \text{ kHz} \\ 2714\text{-}2600 = 114 \text{ kHz} \end{cases}$$

$$b. \begin{cases} 2600 + 57 = 2657 \text{ kHz} \\ 2714 - 57 = 2657 \text{ kHz} \end{cases}$$

$$c. \begin{cases} 2714 + 2(57) = 2828 \text{ kHz} \\ 2(2714) - 2600 = 2828 \text{ kHz} \end{cases}$$

Theoretically, the pairs of responses fall at precisely the same frequency, but the actual tones will be displayed as pairs due to slight input signal deviations in frequency. Figure 5 shows the two 2657 kHz components computed in (b), and Figure 6 shows the output at 2828 kHz.

Figure 6 also illustrates the usefulness of the 7L5's averaging functions to pick a low level signal out of the noise.

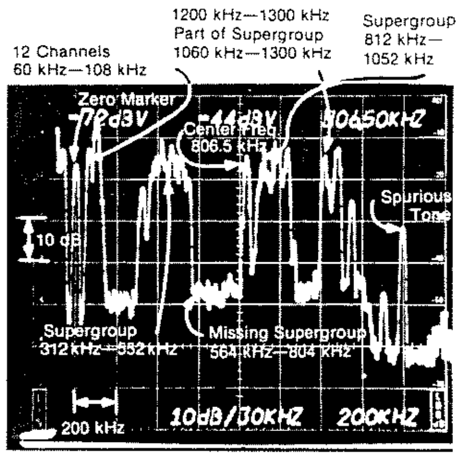


Fig. 3. Full baseband spectrum showing several supergroups.

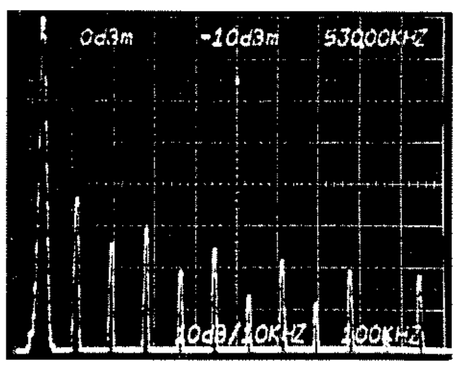


Fig. 4. Display of −10 dBm fundamental and harmonics.

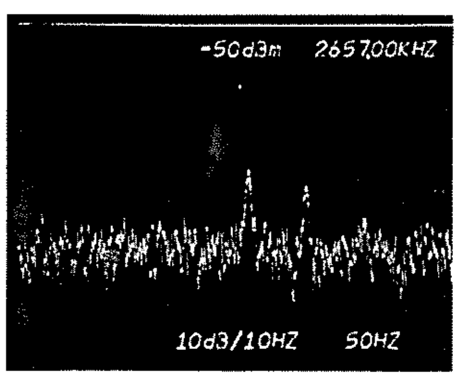


Fig. 5. Three tone intermodulation products

Detailed amplitude analysis

Figure 7 shows a carrier leak at 520 kHz. Carrier leak is the carrier signal that remains after suppression in a suppressed-carrier system. Leakage amplitude is −60 dBm. This is almost 25 dB below the level of the adjacent signals.

Figure 8 shows three channel carriers with the vertical scale expanded by a factor of ten to produce a display factor of 0.2 dB/div. The display was produced by connecting the detected vertical output at the 7L5 front panel to a 7A22 amplifier set at a sensitivity ten times the 50 mV/div output from the 7L5. Accurate vertical calibration is obtained by changing the 7L5 reference level by 1 dB and adjusting the 7A22 variable gain to get the desired vertical deflection (5 divisions in our example). Since the trace now occupies ten screen heights, it is necessary to use the 7A22 dc offset control to position the display on screen. Maximum amplitude difference is 0.6 dB and the minimum difference is 0.15 dB. Considering the close frequency spacing of the signals and the instrument flatness specifications, the measurement has an accuracy of better than 0.1 dB.

Detailed frequency analysis

For detailed frequency observation it is necessary to reduce the resolution bandwidth to the point where individual tones can be resolved. With 10 Hz resolution, the 7L5 can perform a detailed analysis even of frequency-shift-keyed (FSK) signals as illustrated in Figure 9.

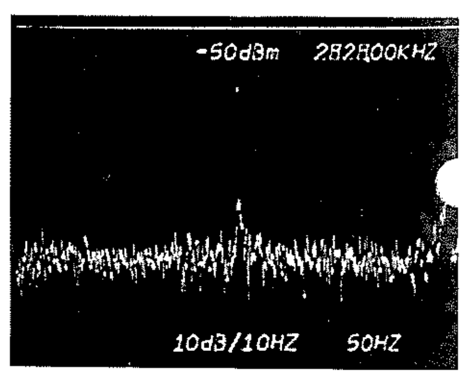


Fig. 6. Three tone intermodulation product at 2828 kHz.

Figure 9 shows the spectral characteristics of the first channel (409 kHz) in group three of the basic supergroup (312-552 kHz). At 50 Hz/div we are only observing 500 Hz of the 4 kHz channel. With individual components resolved, it is clear that this is an FSK signal. Tone amplitude level is −60 dBm (−69 + 9.1)*. At a vertical setting of 10 dB/div we can easily establish the level of the various tones, but the vertical resolution at this setting prevents high accuracy. For high accuracy, it is necessary to go to a vertical setting of 2 dB/div.

Looking at noise

The ability to save, and hence to freeze, a spectral display in the A digital storage section, while section B continues to update, permits some useful measurements. An illustration of the "Save A" feature is shown in Figure 10. The lower trace displays the shape of a notch filter at 2,057 kHz. There are also several spurious signals about 300 kHz below the notch frequency. The upper trace shows the update in Memory B with the MAX HOLD function activated. The MAX HOLD function will hold in memory and display the maximum amplitude level that occurs during observation time. Here a noise burst has increased the system noise by 10 dB while totally obliterating the effect of the notch filter.

Recommended equipment

All of the foregoing measurements can be made using the following equipment. Some of the basic features are listed. Complete specifications can be obtained from your Tektronix Field Engineer or Sales Representative.

7000 Series Oscilloscope: Any 7000 Series Oscilloscope with crt readout will accommodate the 7L5 Spectrum Analyzer Plug-in. The 7603 with its large 6½-inch crt makes an ideal choice.

7L5 Spectrum Analyzer: Plug-in spectrum analyzer with digital storage.

- Frequency range: 20 Hz to 5 MHz. Good for 600 channel systems.
- Resolution: Down to 10 Hz. Will resolve individual FSK tones.
- Sensitivity: Down to −135 dBm. Good for checking those hard to find leaks and spurious tones.
- Frequency accuracy: 5 Hz for accurate tuning to desired channels.

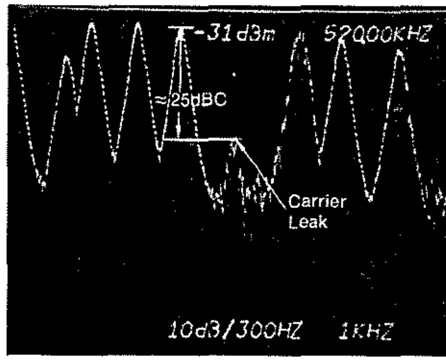


Fig. 7. Carrier leak measurement.

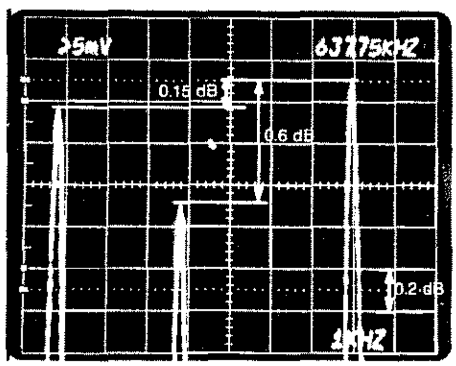


Fig. 8. Expanded amplitude measurement of three channel carriers at 0.2 dB/div.

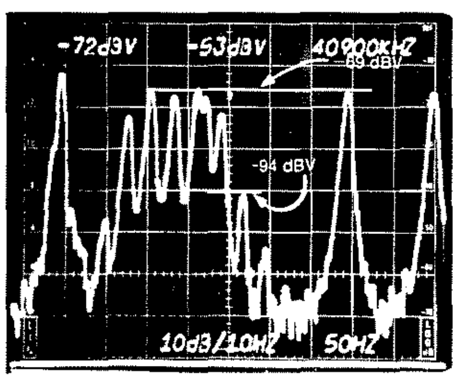


Fig. 9. Spectrum of FSK signal.

L3 Plug-IN Module Option 01: Plug-in front end for 7L5.

- Impedance levels: 1MΩ, 600Ω, 75Ω for bridging and single-ended measurements.

Option 25 Tracking Generator: Provides tracking output level over the full 5 MHz frequency range for checking component transmission characteristics.

013-0182-00 Balanced Input Transformer:

- Frequency range: 50 kHz to 3 MHz, usable from 10 kHz to 20 MHz.
- Built-in terminations: 124Ω, 135Ω and NONE for bridging.

Application notes describing in detail the measurement techniques used here are available on request. Customers in the U.S. and Canada may use the inquiry card in Tekscope. Overseas customers should contact their Tektronix Field Engineer or Sales Representative.

"Baseband Measurements Using the Spectrum Analyzer" — Tektronix AX-3433

"The Spectrum Analyzer as a Frequency Selective Level Meter" — Tektronix AX-3682

"Swept Selective Level Measurements" — Tektronix AX-3861

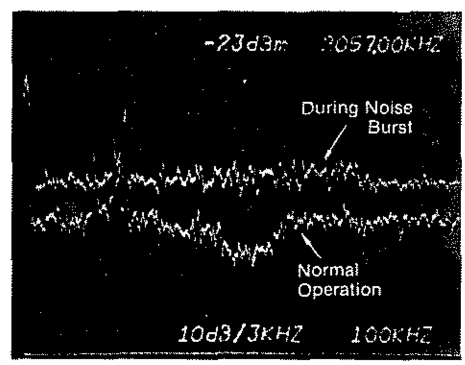
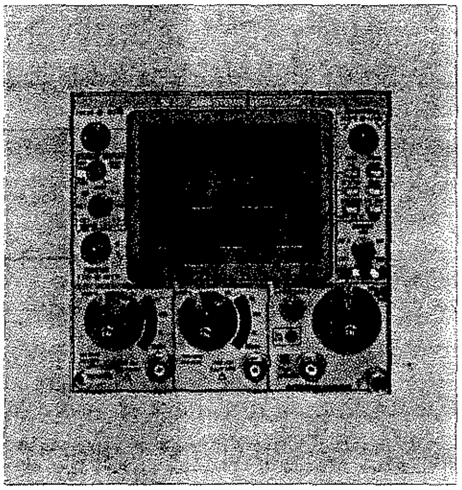


Fig. 10. The "SAVE A" feature allows comparison between normal operation and operation during noise burst.

*At 124Ω, dBm = dBv +9.07

New Products

SC 503 10 MHz Storage Oscilloscope



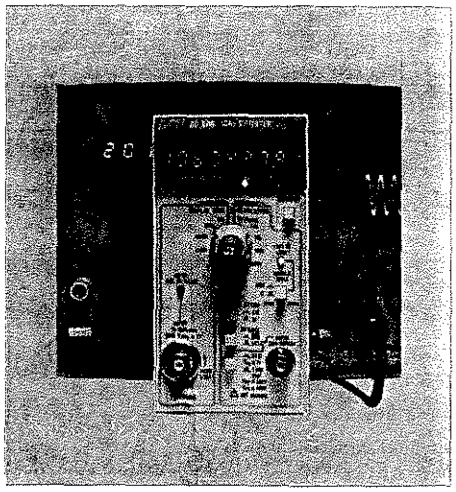
Bistable Storage for TM 500

A new 10 MHz storage oscilloscope is available for the versatile TM 500 Series. Bistable storage with auto erase provides a normal stored writing rate of 50 cm/ms (80 div/ms) which can be enhanced to 250 cm/ms (400 div/ms) by trading off storage time. Maximum storage time is about four hours. In auto erase mode, viewing time can be varied from 1 to 10 seconds. The SC 503 can also be operated in non-storage mode.

Versatile triggering — auto, normal, single-sweep — trigger view, and variable trigger holdoff add to the SC 503's ease of use. Other important features include 3% vertical deflection and time base accuracy. A full range of input modes, including X-Y, is provided.

The SC 503's storage capability makes it ideal for viewing the low repetition rate or low writing rate signals encountered in medical, biophysical, and electromechanical applications.

DC 508 1 GHz Frequency Counter



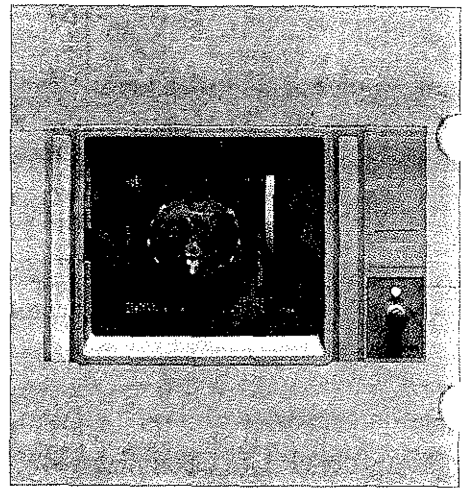
A Gigahertz Counter for TM 500

The DC 508 was designed to extend the TM 500 counter capabilities to include the critical frequency measurement problems that face the communications industry with the opening of the new 800 MHz to 1 GHz band.

The DC 508's 15 mV direct input sensitivity allows transmitter frequency measurements to be made by using a remote whip antenna or an inductive probe. Its X100 audio frequency resolution multiplier permits measuring resolution to 10 milli-Hertz in one second. This feature is particularly useful for communications systems involving frequency shift keying and tone squelch frequencies. For example, a counter without the resolution multiplier will take 100 seconds to get the same resolution.

The DC 508 features a 9-digit LED display which indicates frequencies and totalizes events from 0 to 999,999,999. Frequency range of the DC 508 is 10 Hz to 100 MHz (direct input) and 75 MHz to 1 GHz (prescaler input).

The 634 Video Display Monitor



Three New Monitors Broaden Your Price/Performance Choice

The new 634 Video Display Monitor offers superior resolution, brightness uniformity, and geometry for highest quality performance in medical-imaging applications. Worst-case resolution is 1100 lines (nominally 1400 lines) at a brightness of 100 cd/m² (30 fL). Geometric distortion is 0.5% or less over a 9 cm circle at the center of the screen.

The 624 Display Monitor features improved brightness in a moderately priced X-Y display. A brightness of 130 cd/m² (40 fL) and 12 mil spot size assure crisp, detailed displays even in normal room ambient light. The 10 x 12 cm viewing area permits more data to be displayed more clearly, including complex graphics and alphanumerics.

The new 620 Display Monitor is very lowcost, electrostatically deflected display designed to accommodate OEM's who are now building their own displays. The 620's 10 x 12 cm viewing area, 5 MHz Z-axis bandwidth, 2 MHz vertical and horizontal bandwidth, and 12 kV accelerating potential make it an ideal choice for many price-sensitive, good-performance, OEM applications.

A-4022 7/78

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