touch" automatic preset, and an automatic digital channel indication.

This system first begins a tuning search from the lowest channel on and memorizes the channel number of all available stations step by step in a nonvolatile EAROM, completely automatically. When a channel button is selected, it reads the channel number stored in the EAROM and tunes the selected station through the SAW frequency synthesizing operation.

V. Conclusion

We have successfully developed two TV tuning systems which apply an SAW device in a new way. One is the automatic channel indicating system which uses voltage synthesizing with a channel indication. The other is frequency synthesizing utilizing an SAW comb filter, which is an expanded system derived from the former one. In these systems, a newly developed SAW comb filter is introduced, which has comb peaks at the frequencies where channels are allocated. A channel number is recognized by counting the number of comb peaks which the local oscillator signal goes through. In the SAW frequency synthesizing, a channel is selected by sweeping the local oscillator frequency until the number of comb peaks coincides with the predetermined number according to the channel number.

These systems are also applicable to any system such as NTSC, PAL, and SECAM, and furthermore to CATV receivers.

Acknowledgment

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Implementation of Satellite Communication Systems Using Surface Acoustic Waves

JEANNINE HENAFF AND PIERRE C. BROSSARD

***Abstract*—Current performance of surface-acoustic-wave (SAW) devices offers several advantages in the construction of digital communication networks. Experimental examples of delay lines, filters, oscillators, etc., used for the modulation, the frequency conversion, and the demodulation of *n*-phase-shift-keyed (PSK) digital signals are described and present results are reported. These devices, especially designed for satellite communication systems, operate in the range 70 MHz to 1 GHz where the surface-wave technology allows reduction in size and weight combined with ruggedness and reliability.**

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I. Introduction

THIS PAPER identifies and describes the subunits of equipment involved in a satellite digital communication link where applications of surface-acoustic-wave (SAW) devices can improve significantly their design.

After a brief review of the essential features of the 2- and 4-PSK modulation, the first section deals with the trans-

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mitting end of the radio link. It includes an all-digital modulator which is composed of a 280-MHz SAW oscillator, followed by an intermediate frequency (IF) whitening SAW filter which not only limits the spectral occupancy but also supplies a uniform spectrum at the transmitting end with the constant, flat delay required by the digital signals.

Then, an SAW voltage-controlled oscillator (VCO) is described, which rapidly performs the frequency shifts required to change the carrier frequencies at the transmitting and receiving ends, a necessary requirement in a complex network like a global satellite network. The SAW–VCO is tunable over a ± 3-percent frequency bandwidth around its center frequency, near 1 GHz.

The description of the receiving end, in the second section, is carried out with an IF linear-phase SAW filter limiting the noise bandwidth to the Nyquist bandwidth. The demodulation of the PSK signals is then discussed: differential demodulation using 70-MHz SAW delay lines and coherent demodulation using a 280-MHz (or 210-MHz) voltage-controlled SAW oscillator will be compared. Advantages of coherent demodulation for communication via satellite will be emphasized.

Already, some of these devices have been successfully used for picturephone applications and 30 voice channels–PCM multiplex via Symphonie satellite. For example, two types of differential demodulators using SAW delay lines were experimentally introduced first in October 1975 between Paris-CNET and Genève, Switzerland, for the Telecom 75 Exhibition and later, in November 1976, between the Paris UNESCO Headquarters and the 19th UNESCO General Conference in Nairobi, Kenya [1], [2]. Since 1978, two links have been normally operating between Pleumeur-Bodou (home—France) and La Rivière des Pluies (Island of La Réunion, in the Indian Ocean) via Symphonie or Intelsat 4 for the transmission of the 30 voice channels each. Furthermore, a radio link between Bercenay-en-Othe (France) and La Rivière des Pluies (Island of La Réunion) via Symphonie and/or Intelsat will soon[1] take advantage of most of these SAW devices to transmit 240 voice channels.

II. Short Description of a Satellite Digital Communication Link

A block diagram identifying the subunits of terrestrial equipment involved in a conventional satellite digital communication link is described in Fig. 1. At the transmitting end, we first meet an *interface* (a) between the line carrying the digital information and the actual modulator. It produces 1) the conversion of the HDB3-bipolar signal on line (standard rates in Europe: 2, 8, or 34 Mbit/s) to a binary signal+clock signal; 2) the scrambling, providing the energy dispersal of the radio carrier in order to avoid—in the absence of modulation—the energy concentration on a single frequency [3], [4] and facilitating the bit timing

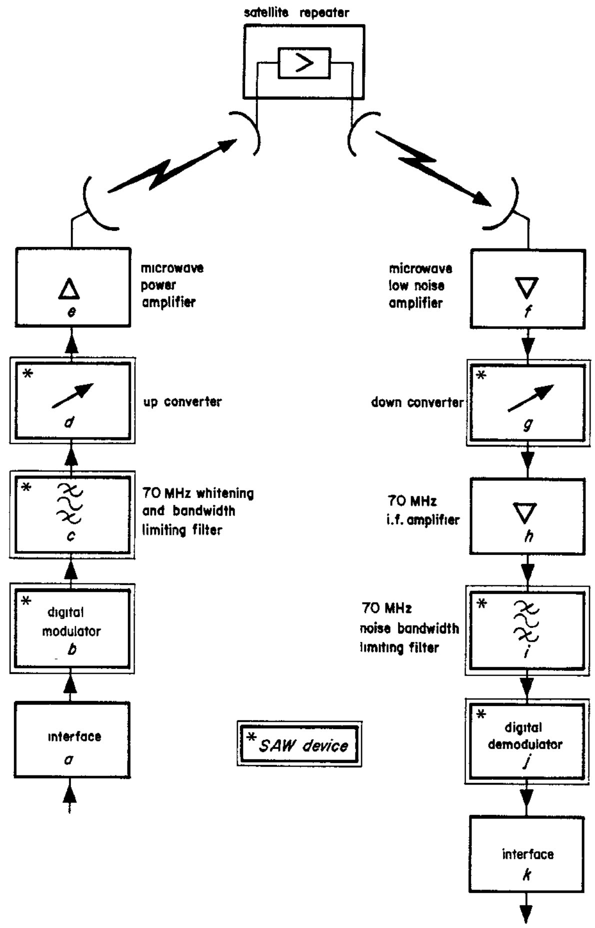


Fig. 1. Block diagram of a satellite digital communication link.

recovery at the receiving end by elimination of long sequences of zeroes; 3) the differential encoding, allowing either the second- or fourth-order ambiguity removal in coherent demodulation of 2-PSK or 4-PSK signals, or a differential demodulation.

Then the signal is processed by a *modulator* (b) where it is two- or four-phase shift keyed at an IF to generate the modulated signal. Due to the differential encoding in the interface a bit "1" is transmitted with a phase shift$=\pi$, a bit "0" is transmitted with a phase shift$=0$, for 2-PSK signals, and for 4-PSK signals, the symbol "11" corresponds to π, "01" corresponds to $\pi/2$, "10" corresponds to $3\pi/2$, "00" corresponds to 0.

The modulator is followed by an *IF filter* (c) limiting the spectral occupancy and "whitening" the spectrum.

The 70-MHz modulated signal is then *upconverted* (d) to the microwave carrier in the 6-GHz range, and a *power amplifier* (e) provides the antenna with the required microwave power.

At the receiving end, the signal coming from the antenna is first amplified in a 4-GHz *parametric amplifier* (f) where it may or may not be cooled, then *downconverted* (g) from the microwave range to the 70-MHz IF, and it is next amplified at the level required at the input of the demodulator in an AGC *70-MHz amplifier* (h).

To limit the noise bandwidth as well as the noise power

[1]In normal operation via Intelsat 4A since September 1980.

collected by the demodulator, an *IF filter* (i) is inserted in front of the *digital demodulator* (j). The demodulation associated with bit timing recovery circuits may be of the differential or coherent type for 2-PSK signals since the bit error rate versus the E/N_0 (energy per bit/noise spectral density) is about the same for the two implementations (for example, the difference on E/N_0 between coherent demodulation and differential encoded demodulation is equal to 0.5 dB for a bit error rate of 10^{-4}), but for 4-PSK signals, the coherent demodulation is obviously preferable and requires a carrier recovery circuit.

Finally, another *interface* (k) between the digital line equipment and the demodulator is needed. It performs the differential decoding in the case of coherent demodulation, and the final conversion of (binary signal + clock) to HDB3 bipolar signal.

SAW devices—delay lines, filters, oscillators, etc.—can improve significantly the design of the following subsystems: (b) modulator; (c) IF transmission filter; (d) and (g) up and down frequency conversion; (i) IF reception filter; and (j) demodulator. Some examples will now be described in a more detailed form.

III. Transmitting End

A. Digital Modulator

Conventionally, at the transmitting end of a 4-PSK modulated link, the input digital signal is split into two bit streams (even and odd pulses) differentially encoded and phase-shift-keyed at the IF to generate the modulated signal (Fig. 2) [5]. As the transmitted digital signal is differentially encoded, it is the phase difference between two adjacent instants t and $t-2T$ (if T is the bit duration) which carries the information. The phase shift 0 or π of the IF represents 0 or 1 of each bit stream; a phase shift of $\pi/2$ in addition to an adder give the 4-PSK IF signal. The conventional device needs two translators to drive the mixers with signals symmetrical in relation to the ground. These translators use wide-bandwidth large-gain dc coupled amplifiers able to deliver a high output level. It is very difficult to build such amplifiers with an offset remaining stable when the temperature varies and, therefore, instabilities can appear especially at high rate.

On the contrary, the ability of SAW oscillators to give a stable frequency directly in the VHF–UHF frequency ranges permits the realization of simple and reliable all-digital modulators [6]. Starting from a 280-MHz SAW oscillator (point A, Fig. 3), a digital divider by 2 gives square waveforms at 140 MHz. Then, two other divisions by 2 of the square waveform (point B) and of its complement (that is to say, with a π phase shift, point C) give the four possible phases of the IF signal on four outputs (points D, E, F, and G). A suitable switching of these four outputs provides the desired phase state according to the binary streams coming from an odd/even splitting and encoding device.

When starting from a 70-MHz bulk quartz oscillator, it is not possible to obtain the four $\pi/2$ phase shifts by using

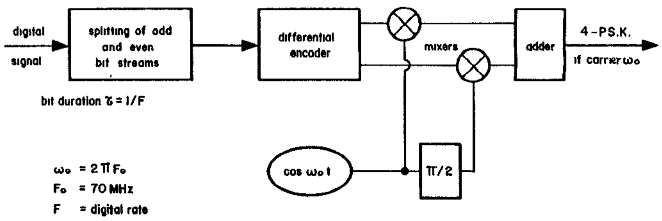


Fig. 2. Block diagram of a differential four-phase PSK modulator.

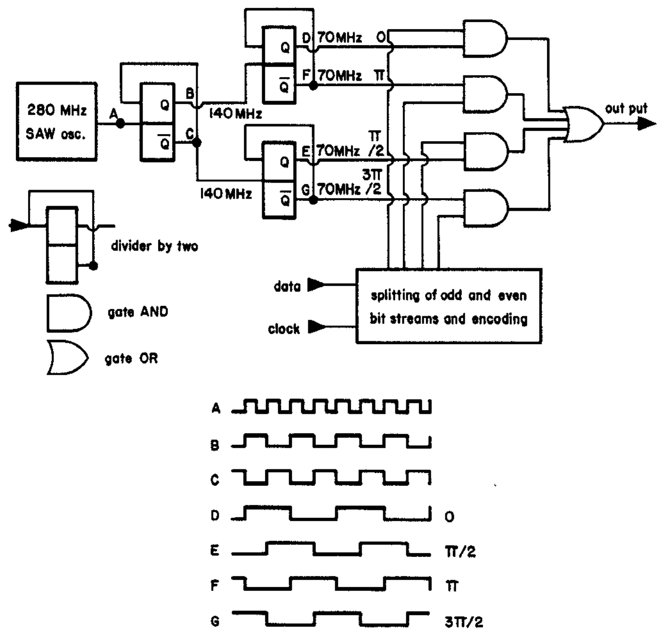
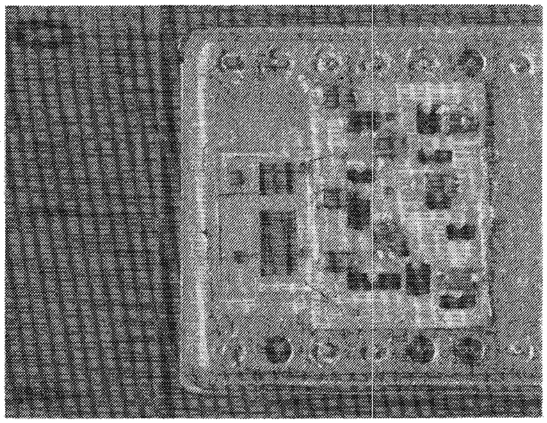


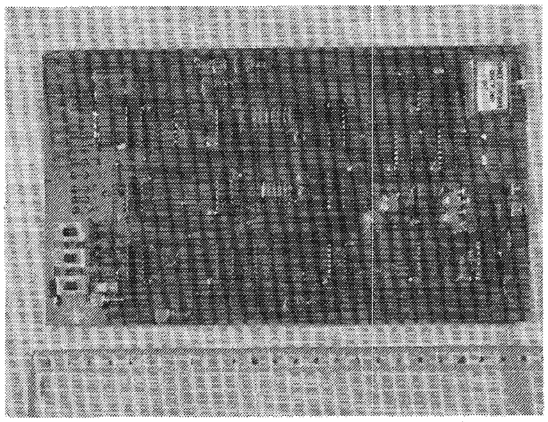
Fig. 3. Block diagram of an all-digital four-phase PSK modulator starting from a 280-MHz SAW oscillator.

only logical circuits, because additional delay lines are needed. Conversely, the division by 4 of a 280-MHz signal from an SAW oscillator gives directly the four phase shifts required. Up to now, the reliability of the SAW oscillators is excellent, and short- and medium-term stabilities are well within the tolerances. For instance, in the frequency domain, the thermal noise floor is exceptionally low: down to -170 dB below the carrier for an oscillator working at 100 MHz and the frequency varies by 4 ppm in the temperature range 10°–40°C when using an ST-cut quartz crystal as substrate. As far as the long-term behavior is concerned, the experimental results are not yet very significant as these oscillators are relatively recent. Nevertheless, measurements performed in various laboratories agree in predicting an aging rate of about 1 ppm/yr at the present state of the art [7], [8].

Fig. 4(a) shows a photograph of the SAW oscillator using a 280-MHz SAW delay line designed on a 4×8 mm ST-cut quartz, ultrasonically bonded to the active circuit printed on an alumina substrate. The whole device is implemented using only "off-the-shelf" logical circuits and without any analog circuit liable to drift (cf. Fig. 4(b)). The four phase states are keyed at $\pm 1°$ and the levels are equal to within ± 0.5 dB. These results are about the same as for analog modulators but the temperature behavior and the



(a)



(b)

Fig. 4. Digital modulator. (a) Photograph of the 280-MHz SAW oscillator #OSF 280. (b) Printed board showing the device.

aging of this digital modulator are distinctly better.

An all-digital 2-PSK modulator is obviously easier to implement since only one divider-by-two (to bring down the frequency from 140 to 70 MHz) and half the switches are required.

B. *Transmitting Filter* [9]

In digital transmission, the Nyquist criteria are generally expressed in terms of signaling with impulses [10]. This is particularly convenient because the spectrum of an impulse is a constant at all frequencies (white spectrum). If any other pulse shape is used, the criteria are satisfied by inserting a network which changes the given pulse spectrum to a constant [10], [11]: this is sometimes termed a "whitening filter." In fact, we use rectangular pulses and, therefore, the density of an n-PSK signal is $(\sin \pi FT/\pi FT)^2$ distributed, where F is the frequency and T is equal to the symbol length; that is, 1 bit for 2-PSK signals and 2 bits for 4-PSK signals.

As the first Nyquist criterion restricts the required bandwidth to $F_0 \pm 1/2T$, the spectral occupancy can be limited by filtering the IF PSK signal. Moreover, it is possible to correct the transmitting filter, giving it a $(\pi FT/\sin \pi FT)$ amplitude response to obtain a white spectrum at the transmitting end. This "double-humped" filter is then a whitening filter [12].

SAW bandpass filters can very easily perform this function [13]. Moreover, they can be designed with a linear phase response (symmetric finite impulse response) and thus have constant group delay. The synthesis of such a frequency response, or rather its low-pass prototype, may be achieved by computing the corresponding impulse response [14]. In fact, we require only correct behavior in the frequency range $F_0 \pm B$ (where F_0 is the central frequency and $2B$ the 3-dB bandwidth); then the prototype frequency response will be

$$s(F)=\begin{cases} \dfrac{\pi FT}{\sin \pi FT}, & \text{for } |F| \leqslant B \\ 0, & \text{for } |F| > B. \end{cases}$$

The inverse Fourier transform of the response is given by the convolution theorem when rewriting the above formula as

$$s(F)=R_B(2\pi F)\times(j\pi FT)\times vp\frac{1}{j\sin(\pi FT)}$$

where $R_B(2\pi F)$ is the rectangular pulse function of width $2B$.

We have next [15]

$$h(t)=\frac{\sin \pi Bt}{t} * \frac{1}{2}\delta' * \sum_{k=0}^{\infty} \delta[t+(2k+1)T] - \delta[(t-(2k+1)T]$$

where δ is the Dirac function, δ' its derivative, and where $*$ stands for convolution. Finally, the relevant infinite impulse response is

$$h(t)=\sum_{k=0}^{\infty} \phi[t+(2k+1)T]-\phi[t-(2k+1)T]$$

where

$$\phi(t)=\delta' * \frac{\sin \pi Bt}{2t}=\frac{\pi B}{2}\left(\frac{x\cos x-\sin x}{x^2}\right)_{x=\pi Bt}.$$

Therefore, we have to use a Hamming, or, better still, a Dolph–Chebyshev window, to recover a finite impulse response. Several such filters for 2-PSK and 4-PSK signals with in-line rates ranging from 2.048 to 16.896 Mbit/s, have been designed around the 70-MHz IF.

The example described in Fig. 5 corresponds to a satellite digital communication link (120 voice channel PCM multiplex, 4-PSK modulation) protected by a convolutional code of a ratio equal to 5/4. Curve *a* shows the spectral density of the 4-PSK modulated signal at the output of the modulator. Owing to the use of rectangular pulses, the spectrum distribution is not white but $(\sin \pi FT/\pi FT)^2$. It is, therefore, possible to compute the optimum amplitude against frequency response of the transmit filter (curve *b*), compensating for this decrease and limiting the spectral occupancy to the Nyquist bandwidth. This ideal filter is not easy to realize since its

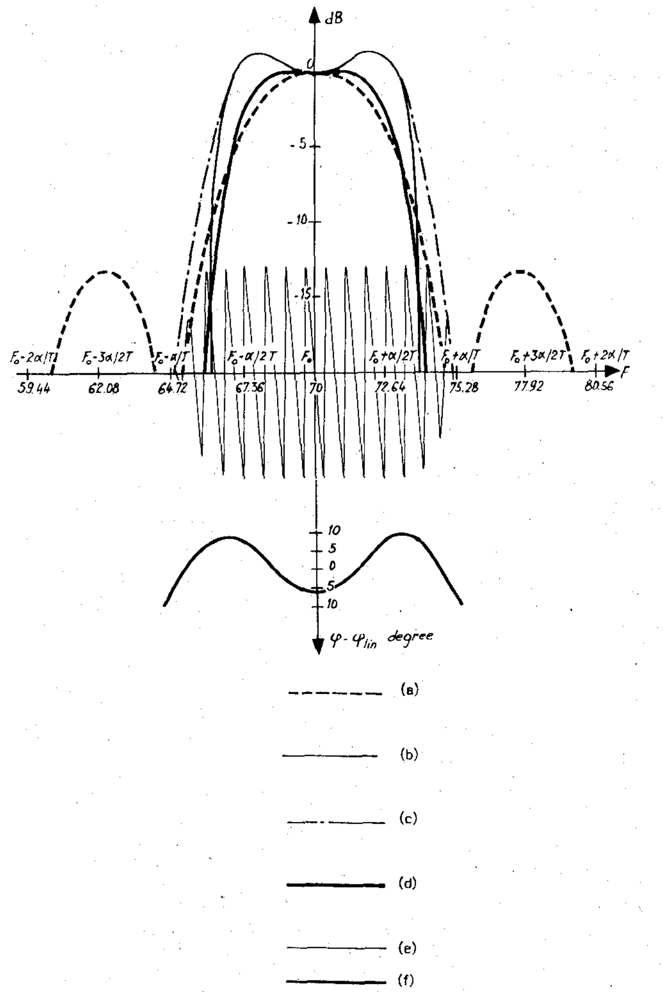


Fig. 5. *a*—Spectral density of the four-phase PSK-modulated signal at the output of the modulator. *b*—Optimum amplitude against frequency response of the transmit filter. *c*—Actual amplitude against frequency response of the SAW transmit filter. *d*—Spectral density of the transmitted four-phase PSK signal. *e*—Experimental phase variation (total). *f*—Experimental phase variation: departure from linearity.

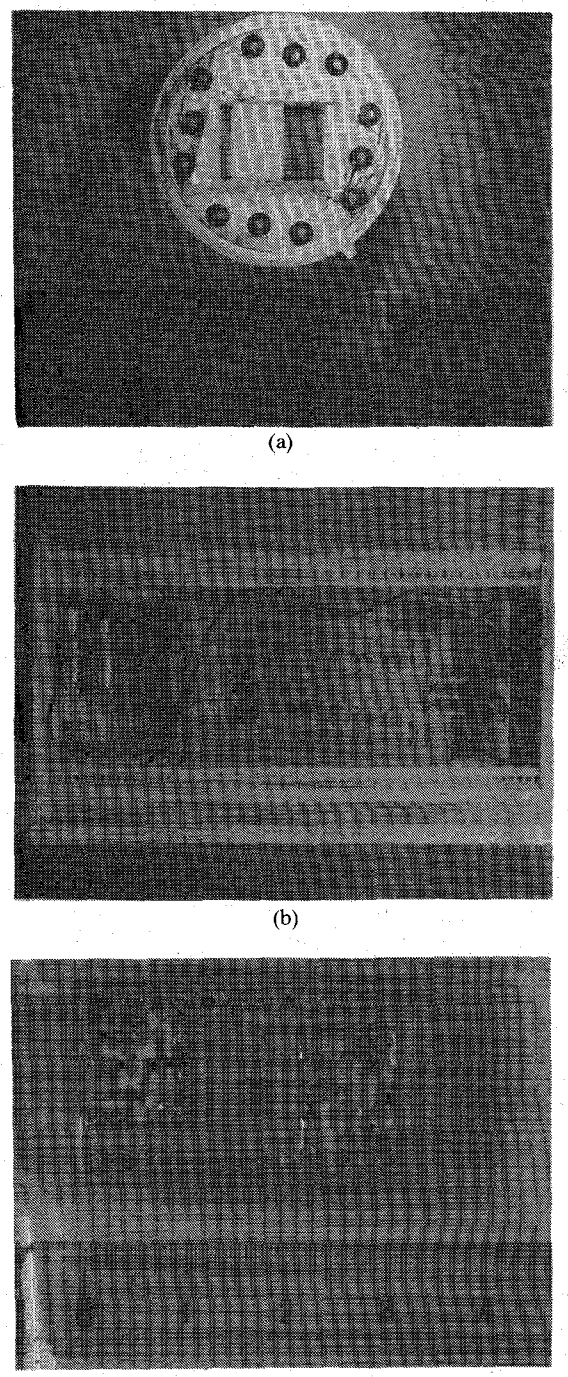


Fig. 6. Whitening filter. (a) Photograph of the SAW linear phase 70-MHz whitening filter. (b) Printed board including two such filters. (c) Photograph of an integrated version of an SAW whitening filter between two hybrid amplifiers.

steepness is quasi-infinite, and so an SAW transmit filter with an actual amplitude against frequency response, as shown in curve *c* of Fig. 5, has been implemented on a 5×8 mm, 0.5-mm thick *YZ* lithium tantalate substrate (cf. Fig. 6(a)). It works at 70-MHz center frequency (wavelength $= 46.3$ μm) with a transducer aperture of 2.5 mm. The regular transducer consists of 13 finger pairs and the second one is a 100 weights apodized transducer, leading to an $x/\sin x$ frequency response in the bandpass. The insertion loss of the filter itself is equal to 27 dB, but it is inserted between two hybrid integrated amplifiers which provide good matching and the required level. The printed circuit including these devices is shown in Fig. 6(b) and Fig. 6(c) shows a photograph of an integrated version of another similar filter. The 1.5-dB humps and 9-percent 3-dB bandwidth are easily obtained and the 20-dB bandwidth has been increased up to 13 percent. The resulting spectral density of the transmitted 4-PSK signal is displayed in Fig. 5*d*; it is a good compromise between a white spectrum and a Nyquist bandwidth limitation. The experimental phase ϕ showing a linear variation is plotted in Fig. 5*e*, and the departure from linearity in Fig. 5*f*. The maximum deviation is equal $\pm 8°$, well within the limits of tolerance. The group delay is then constant and equal to 1.43 μs. No equalizer or subsequent adjustment will then be needed.

The experimental and theoretical bit error rate (BER) against energy per bit/noise spectral density (E/N_0) are

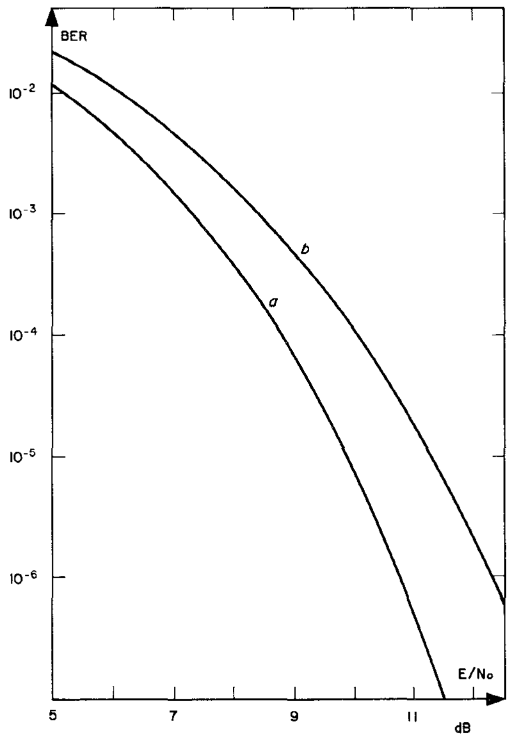


Fig. 7. Bit-error rate (BER) versus (energy per bit/noise spectral density)(E/N_0). Four-phase PSK: differential encoding and coherent demodulation. *a*—theory, *b*—experiment using transmitting and receiving SAW filters.

shown on Fig. 7 for a complete transmit–receive unit using such a $(\pi FT/\sin \pi FT)^2$ filter at the transmitted end and a conventional SAW filter at the receiver end. Notice that the discrepancy between the theoretical and experimental results is only equal to 1 dB, which is considered exceptionally good.

C. *Upconverter*

A frequency shift, using a transmitted local oscillator and a mixer, is necessary to convert the 70-MHz modulated signal to the microwave signal sent to the antenna. In the fixed-frequency radio links, this local oscillator is generally obtained starting from a bulk quartz oscillator followed by successive amplifications and multiplications. However, the satellite system operators have to change their transmission frequency according to the satellite in operation, the digital rate, the allocated repeater, the location of their carrier within the frequency plan, etc.

To avoid ceaseless changing of the quartz oscillator and readjustement of the multipliers, the operators prefer the use of a frequency synthesizer as primary pilot, which gives the so-called "agile" frequency shifts. For this purpose, SAW devices provide a very simple and efficient solution since voltage-controlled stable oscillators are obtained which can, however, be tuned over a sufficient frequency range. Moreover, they oscillate directly on the fundamental in the UHF range. Thus the multiplication factor is low and the filtering of spurious spectral lines becomes easier.

As an example, one can describe the upconverter for a 6–4-GHz satellite system: in the earth station, the transmit frequency is in the range 6.2 ± 0.2 GHz, and, therefore, the pilot frequency has to be tuned over a $\pm 0.2/6.2 = \pm 3.2$-percent range. This can be done using a 1.033-GHz SAW oscillator voltage controlled to give the ± 33-MHz tuning range and followed by a multiplier by 6. The transmitted local-oscillator arrangement is shown on Fig. 8. The wanted frequency is manually selected between 1001.000 and 1065.000 MHz. The first 1 and 0 are fixed and, therefore, the desired frequency can be digitally expressed with only 16 bits. This frequency is compared with the measurement of the frequency of the SAW oscillator and the dc difference signal is used to control the electronic phase shifter of the voltage-controlled SAW oscillator.

The voltage-controlled SAW oscillator is especially designed for low-Q, large tuning capability. Fig. 9(a) presents a photograph of the UHF hybrid integrated amplifiers and, on the left part, the 2×3-mm SAW delay line. In the present experiment, the two transducers of the SAW delay line are placed edge to edge, the distance between the transducers being equal to their length (inset of Fig. 9(a)). They are 11 fingers—3.3-μm wavelength interdigital transducers of 150-μm aperture. The ± 33-MHz tuning frequency range is covered for ± 2-V variation applied to the voltage-controlled phase shifter using two NEC 1S1619 varactors. Another photograph of the device, Fig. 9(b), shows the electronic phase shifter with its hybrid quadrature junction ANAREN type 10264-3.

This oscillator has to be compared to a conventional

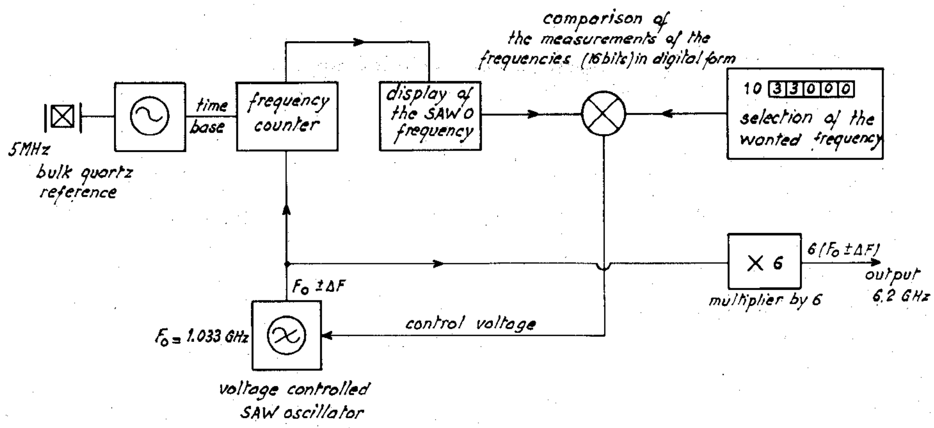
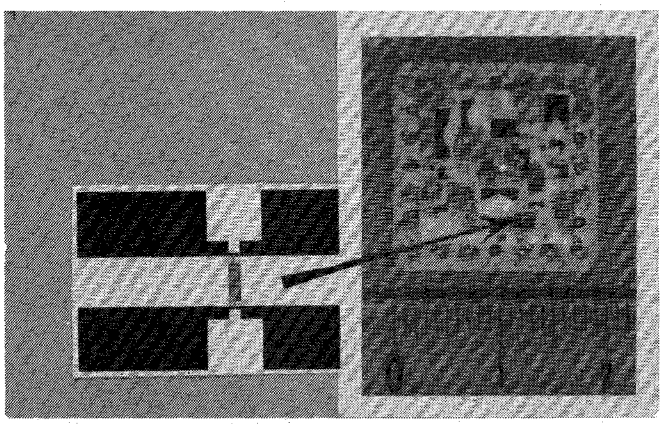
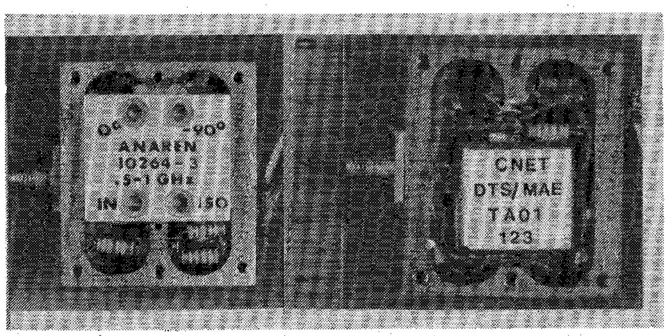


Fig. 8. Upconverter: block-diagram of the local oscillator.



(a)



(b)

Fig. 9. Voltage-controlled SAW oscillator. (a) Main loop of the oscillator with the 1.033-GHz SAW delay line on the left (enlarged view in the inset). (b) Two views of the SAW VCO showing the oscillator and the electronic phase shifter.

VCO or to a VCXO. The frequency range is easily covered with a conventional VCO but the phase noise is not sufficient due to the low quality factor of the VCO. On the contrary, the VCXO has an excellent phase noise but it is not possible to cover the whole frequency range (± 3.2 percent) with a single phase-locked loop (PLL). The voltage-controlled SAW oscillator provides a middle solution: the phase noise, though not as good as with a VCXO, is satisfactory, since the BER degradation is not measurable, and a single PLL allows the coverage of the frequency range. This gives a significant advantage in terms of cost over a conventional synthesizer. Moreover, due to the small hybrid integrated circuits and the low dissipated power, reductions in size and weight are possible.

IV. Receiving End

A. Downconverter

The advantages and the block diagram of the reception downconverter are obviously about the same as those of the transmit upconverter. Describing the same example as in Section III, the down link frequency is in the range of 4

GHz, the multiplier by 6 has simply to be replaced by a multiplier by 4 if the tuning range of the pilot is increased to $\pm 0.2/4 = \pm 5$ percent, which is a little difficult to realize. Another possible solution is to keep the multiplier of 6 and choose, for the voltage-controlled SAW oscillator, a frequency around 0.67 GHz which is easier to implement but leads to a light extra-degradation of the phase noise ($\simeq 3$ dB). However, definitive experimental results are not yet available at CNET.

B. IF Receiving Filter

Even though, at the reception end, "double-humped" bandpass filters are not required since the need is only to limit the noise bandwidth to the Nyquist bandwidth, the usual advantages of the SAW filters remain still important: easy reproducibility, long-term stability (no adjustement), narrow-bandwidth design for low digital rates, and especially linearity of the phase versus frequency which excludes the expensive and difficult to adjust group delay equalizer.

C. Digital Demodulator

Beside the previous examples of SAW applications, several very interesting solutions of digital demodulation by means of SAW devices will now be discussed.

Demodulation of n-PSK signals is achieved either by comparison with the preceding bit, which is called differential demodulation, or, in a more optimal way, through coherent demodulation which requires generating a local reference wave in phase with the carrier (or with the signal shifted to the intermediate frequency) without modulation.

The voltage-controlled SAW oscillators are particularly advantageous for producing such a reference wave thanks to their high oscillation frequency (SAW oscillators offer good performances at frequencies ranging from about 100 MHz to 1 or 2 GHz) and to the broad tuning frequency range covered. As a matter of fact, the Q factor of an SAW oscillator is a design parameter which may be chosen according to very flexible specifications. For the implementation of VCO it is possible to design the acoustic path of the SAW delay line to obtain a tuning range within a given bandwidth, generally 10^{-3} to 10^{-2} but extensions towards some 10^{-2} or 10^{-4} are quite as much realizable.

1) Differential Demodulation [*1*], [*2*], [*5*]: Let us first consider the case of a 2-PSK signal, generally used for low digital rates (2.048 Mbit/s) because its spectral occupancy is twice as much as for 4-PSK signals. Differential demodulation is commonly used for 2-PSK signals because of its very simple, and hence reliable, implementation and also because of the low loss of performance between differential and coherent demodulation in this case (0.5 dB).

Fig. 10(a) shows the block diagram of a 2-PSK differential demodulator: a phase discriminator receives on one input the modulated signal and on the second one the modulated signal delayed approximately by 1-bit duration ($T = 1/2.048 \times 10^6$) by a delay line τ; therefore, its output

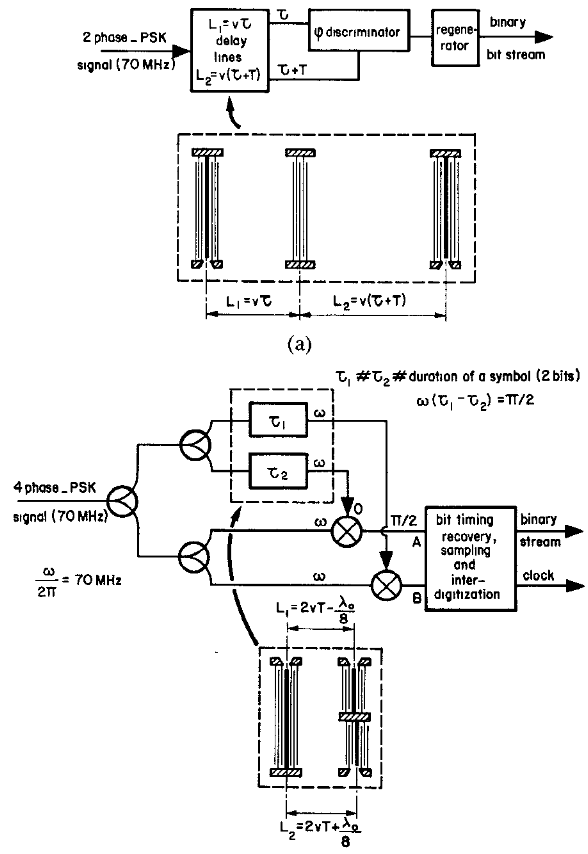


Fig. 10. n-phase PSK differential demodulators. (a) Two-phase PSK demodulator and SAW implementation of the delays. (b) Four-phase PSK demodulator and SAW implementation of the delays.

will provide the original bit stream after regeneration. The delay τ has to be also an integral number of half periods of the 70 MHz IF. Thus we chose $\tau = 68(10^{-6}/2 \times 70) = T(1 - 5 \times 10^{-3}) = 486$ ns.

For high-rate communication systems, the demodulator delay lines are easily realized with coaxial cables; but here for lower bit rates, this approach is no longer valid due to the length ($\simeq 100$ m) and the attenuation of the required cables. An attractive solution to this problem is provided by SAW delay lines and Fig. 10(a) shows also the implementation of the delays. The device includes a bidirectional input transducer, 11 fingers, 7-mm width, and two unidirectional 15 fingers, 7-mm width, output transducers. The distances between the middle of the input transducer and the two output transducers are $L_1 = 2.5$ mm and $L_2 = 4.2$ mm, and when correctly matched, the delay lines have an insertion loss of 10 dB. The piezoelectric substrate is a 9×9-mm YZ lithium niobate, 1 mm thick, leading to a low-cost small-size and easy-to-adjust device.

Differential demodulation of 4-PSK signals can be achieved in the same way. The block diagram of a 4-PSK differential demodulator and SAW implementation of the delays is shown on Fig. 10(b). The device includes a unidirectional input transducer, 25 fingers, 7-mm width, and two unidirectional 25 fingers, 3-mm width, output

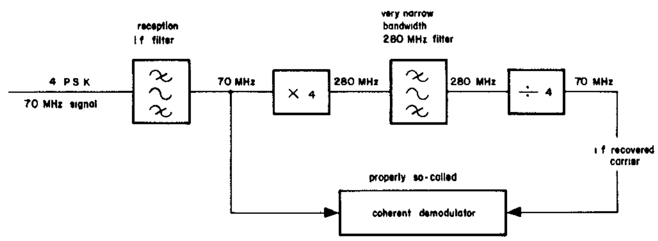


Fig. 11. Simple IF carrier recovery circuit.

transducers. The center frequency is $F_0 = 70$ MHz and the distances between the middle of the input transducer and the two output transducers are: $L_1 = 2vT - \lambda_0/8$ and $L_2 = 2vT + \lambda_0/8$. The piezoelectric substrate is YZ lithium niobate, and, when correctly matched, the delay lines have an insertion loss of 7 dB and a 3-dB bandwidth of 3 MHz, for minimum requisite bandwidth of 1.35 MHz. A very simple, cheap, and reliable solution may still be obtained. But in order to get the same BER, a 2-dB increase of the E/N_0 ratio (energy per bit/noise spectral density) must be available with differential demodulation with regard to coherent demodulation.

Until now, all the differential demodulators realized at CNET operate at a digital rate of 2.048 Mbit/s with an IF of 70 MHz, but this IF can be chosen between 10 and some 100 MHz and the data rates between 100 kbit/s and 10–20 Mbit/s.

2) Coherent Demodulation [*16*], [*17*]: For coherent demodulation, mainly used for 4-PSK signals, a carrier recovery circuit is necessary. The 70-MHz IF is first multiplied by 4 to eliminate any modulation by changing the phase back to $2K\pi$, whatever the original phase shift. After filtering, the 70-MHz recovered carrier is obtained through division by 4 by means of low-cost ECL logic circuits which requires no adjustment (cf. Fig. 11).

When geostationary ultrastable satellites will be available, that is to say if Doppler effect and drifts of the frequency shift in the satellite were negligible, a usual filtering of the 280-MHz spectral line will be sufficient. Then, a solution—very difficult to implement in any other way—may be given by ST-quartz SAW filters thanks to their ability to provide a very-narrow-bandwidth 280-MHz filter with a minimum phase design and thus a low group delay, stable and reproducible [18], [19].

As a matter of fact, frequency shift in the satellite may result in an important variation of the IF in the repeater, and so a PLL is generally preferred for the filtering of the 280-MHz spectral line. This gives narrower filtering and wider tracking range of the intermediate carrier frequency at the price of a larger response time. Then the conventional block diagram would be as shown on Fig. 12(a).

Again SAW devices bring an important advantage. Analog multipliers are devices calling for filtering of the required harmonic frequency, whereas frequency dividers can be fully digitized and are, therefore, inexpensive and more reliable. Since it is very easy to realize voltage-controlled SAW oscillators working directly on the fundamental at

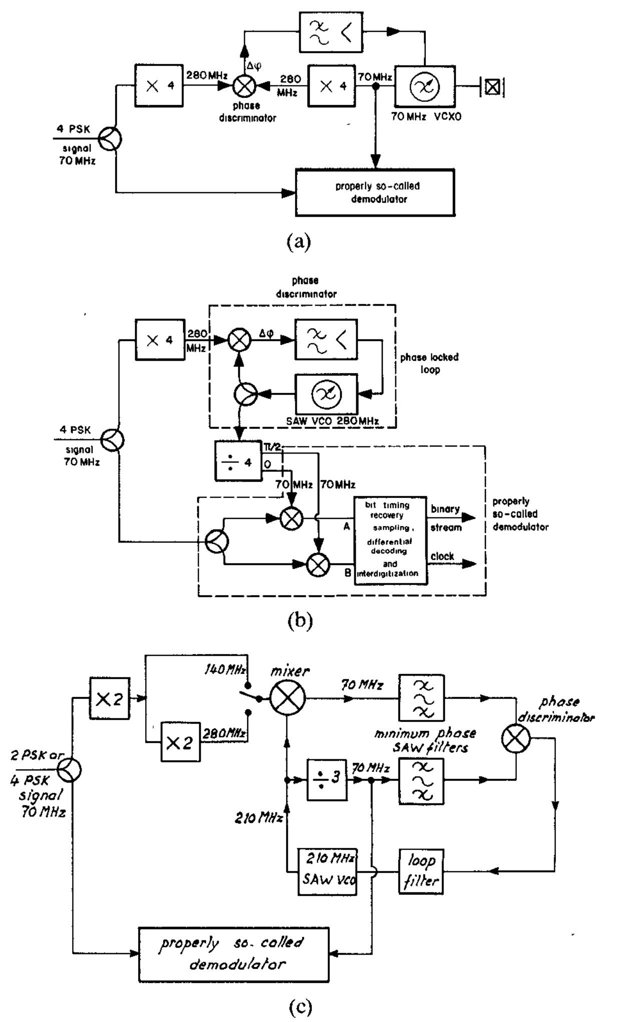


Fig. 12. IF carrier recovery circuits using a PLL. (a) Conventional. (b) Using a 280-MHz SAW VCO. (c) Using a 210-MHz SAW VCO (2-PSK or 4-PSK signals).

280 MHz (cf. Section III-A), the block diagram of the IF carrier recovery circuit becomes the one that is shown on Fig. 12(b) using a divider by 4 instead of a multiplier by 4.

Moreover, an IF carrier recovery circuit operating for 2-PSK *and* 4-PSK 70-MHz signals can be easily implemented [20]. Such a recovery circuit is shown in Fig. 12(c), using a 210-MHz SAW VCO. For 2-PSK signals, π phase shifts are eliminated using a multiplication by 2, and in the same manner for 4-PSK signals, $\pi/2$ phase shifts are eliminated using a multiplication by 4. The nonmodulated signal obtained in this way at 140 MHz (for 2-PSK) or 280 MHz (for 4-PSK) is mixed with the output signal from the SAW VCO working at 210 MHz, and, therefore, a 70-MHz signal is always achieved. This 70-MHz frequency is filtered in a narrow-bandwidth SAW filter and phase compared with the 70-MHz signal resulting from the division by 3 of the SAW VCO frequency (after going through a 70-MHz SAW filter providing the same delay as the first SAW filter).

Due to the good reproducibility of SAW filters it is very easy to obtain exactly the same group delay for the two

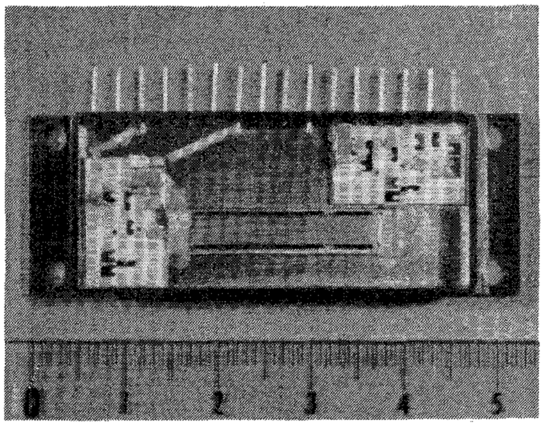


Fig. 13. Photograph of the minimum phase 70-MHz SAW filter #FRP06.

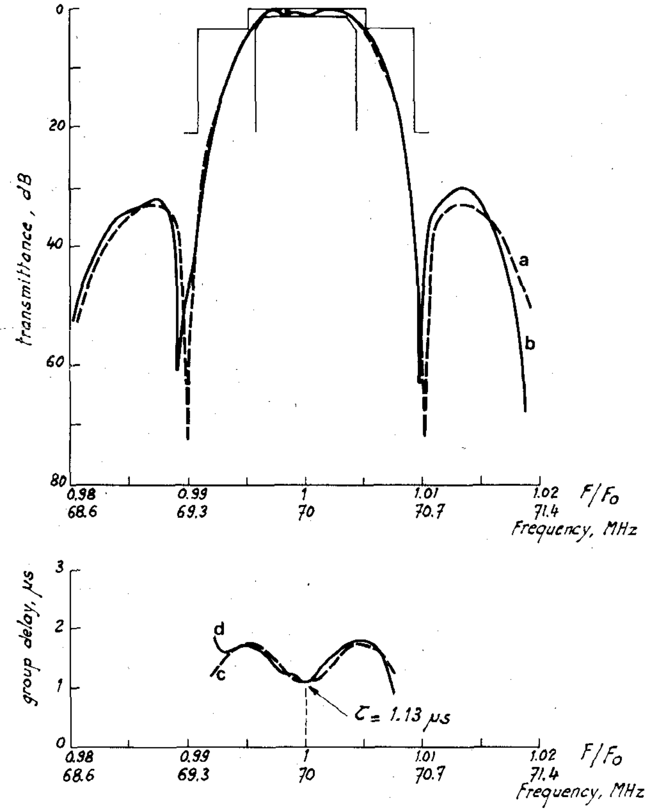
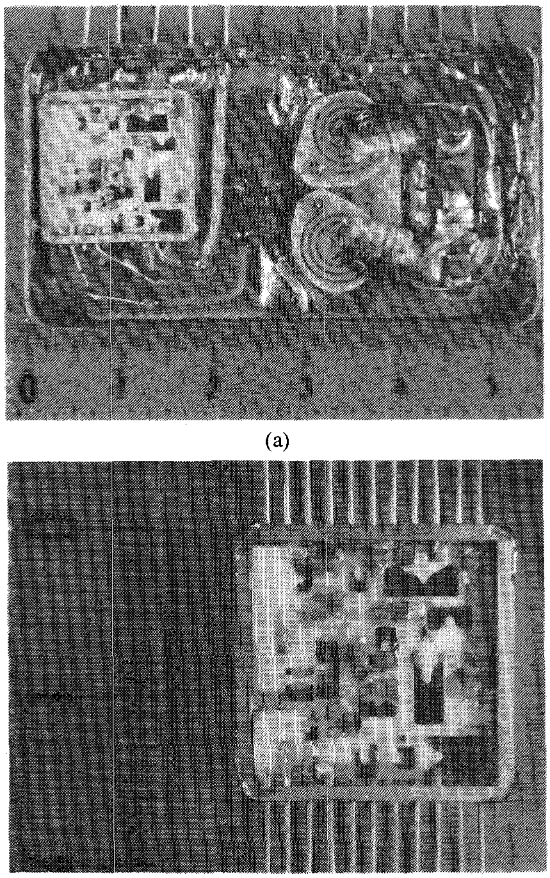


Fig. 14. FRP06 filter frequency responses. a—computed transmittance, b—experimental transmittance (10 dB/div), c—computed group delay, d—experimental group delay (1 μs/div).

70-MHz filters. Fig. 13 shows a photograph of the 70-MHz SAW filter #FRP06. As shown on Fig. 14 a, b, the 1-dB bandwidth is equal to 500 kHz, the 3-dB bandwidth to 700 kHz, and the 20-dB bandwidth to 980 kHz. To obtain such a steepness with a linear-phase filter, a group delay of about 6 μs is necessary. But, when the loop delay is long, the lock-in range of the PLL is small. To reduce the group delay, the SAW filter was then synthesized as a minimum phase filter. The theoretical and experimental group delays are shown in Fig. 14c, d and the minimum value is equal to 1.2 μs only. This filter has been implemented on a 25×5-



(a)

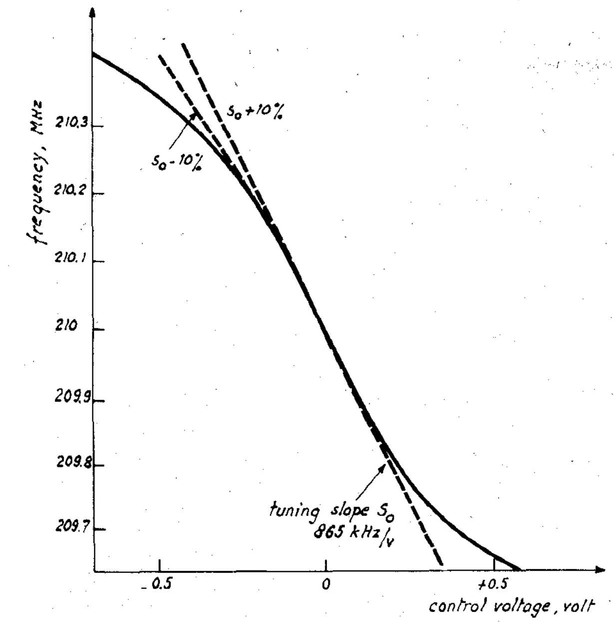
(b)

Fig. 15. 210-MHz SAW VCO pictures. (a) Photograph of the device #OSVC8D. (b) Main loop of the #OSVC8D with the 210-MHz delay line.

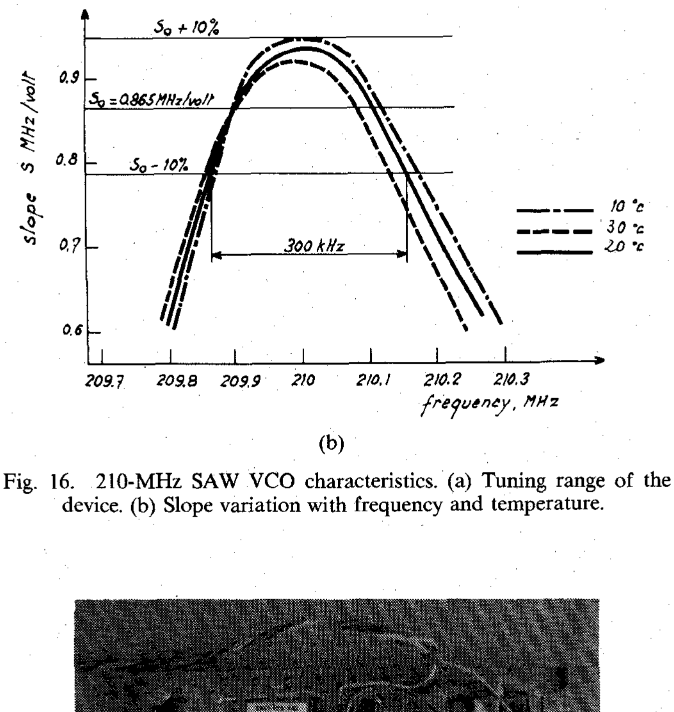
mm ST-cut quartz with a transducer aperture of 3 mm. The two transducers are contiguous with a ground line to separate the input and the output transducers. The regular transducer consists of 100 finger pairs and the second one is a 648 weights apodized transducer. The insertion loss of the filter is equal to 26 dB, and, therefore, the filter is inserted between two hybrid integrated amplifiers which provide the required level (Fig. 13).

Fig. 15 presents a photograph of the 210-MHz SAW VCO. The voltage-controlled phase shifter is seen on the right side of Fig. 15(a) and the 210-MHz SAW delay line integrated on the alumina substrate on which the active circuit is printed is shown on the left side of Fig. 15(a) and in an enlarged view in Fig. 15(b). The tuning range of the device is shown in Fig. 16(a): more than 600 kHz of electronic tuning is achieved with a variation of less than 1 V following an average tuning slope of 865 kHz/V. The variations of this slope with temperature and frequency are given in Fig. 16(b): ±10-percent slope variations lead to ±150-kHz frequency range. Fig. 17 shows a photograph of the IF carrier recovery circuit where the 210-MHz SAW VCO and the two #FRP06 filters of the PLL are clearly seen.

Such a PLL provides very interesting results: 1) ±60-kHz lock-in range for a 0-dB E/N_0 ratio, the noise power being applied before the signal, and more than ±200-kHz



(a)



(b)

Fig. 16. 210-MHz SAW VCO characteristics. (a) Tuning range of the device. (b) Slope variation with frequency and temperature.

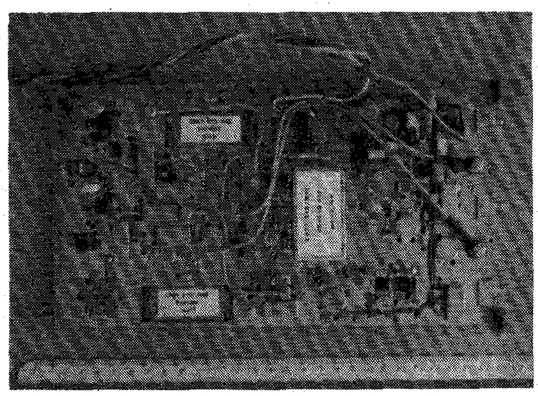


Fig. 17. Coherent demodulator: printed board of the IF carrier recovery circuit showing the 210-MHz SAW VCO and the two FRP06 filters of the PLL.

hold-in range; 2) $<\pm 2°$ residual phase error of the recovered carrier in the 70-MHz$\pm$60-kHz range; 3) 2.5° mean phase deviation for a 0-dB E/N_0 ratio; and 4) less than 1 cycle slipping of the recovered carrier a day, always for a 0-dB E/N_0 ratio.

Such characteristics are obviously very important for a satellite digital communication link protected by a convolutional code and using a Viterbi algorithm decoder [21]. (Remember that a 0-dB E/N_0 ratio will give only a theoretical BER equal to 0.15 when using coherent demodulation of 4-PSK differential encoded signal.) Recourse to analog circuits like demodulation–remodulation devices or Costas loops can then be avoided. Carrier recovery using multiplication by 2 (or by 4) of the PSK-modulated signal or carrier recovery by means of a Costas loop are indeed two ways leading to the same result. But, it is easier to perform the multiplication of the modulated signal using 70-MHz frequency doublers, commercially available, than to use baseband multipliers for the Costas loop. Notice that the VCO of the Costas loop should be also an SAW VCO operating at 70 MHz, with the advantage, over a bulk VCO, of a wider tuning range. At last, the switching from 2-PSK signals to 4-PSK signals is simpler using the circuit described in Fig. 12(c) than for a Costas loop.

V. Conclusion

Implementation of several SAW Devices used in satellite digital communication link has been described in this paper. Table I summarizes the main SAW applications in French satellite systems.

The first experiments using SAW differential demodulators allowed picturephone transmission (video-conference) during exhibitions or conferences and next, picturephone or 30 PCM voice channels plus 6 telegraphic channels transmission via the Symphonie satellite. The experimental links included the Symphonie earth station in Pleumeur Bodou, with its 16.50-m dish, and the small mobile station SAMSON, with its 4.8-m dish.

Since 1978, 2×30 PCM voice channels multiplex, still sing SAW differential demodulators, have been regularly run between Pleumeur-Bodou (France) and La Rivière des Pluies (Island of La Réunion) via Symphonie or Intelsat 4A. In July 1980, a test link was installed between Pleumeur-Bodou and La Rivière des Pluies via Symphonie using SAW devices in the digital modulator, the transmit and receive IF filters, and the coherent demodulator, to transmit 240 PCM voice channels. Definitive running of the 240 PCM voice channels multiplex took place in September 1980 between another French satellite earth station, Bercenay-en-Othe and La Rivière des Pluies via Intelsat 4A.

At present, any SAW device is tested in flight applications, but CNET is making future plans for using SAW filters and demodulators in the satellite itself, if the good results already obtained for ground equipment—especially the reliability—are confirmed.

In conclusion, SAW techniques have proved to be a fruitful solution for extending the use of differential demodulation at low digital rates, and for improving the design of coherent demodulation. They provide whitening or noise-bandwidth-limiting linear phase IF filters but also

TABLE I
SOME SAW APPLICATIONS IN FRENCH SATELLITE SYSTEMS

Date	Object	Link Ends; Data		Satellite	SAW Devices
June 1975	Spacial exhibition Le Bourget (France)	Paris-Le Bourget (mobile station SAMSON)	Pleumeur-Bodou (France) (satellite earth station)	Symphonie	
		Picturephone			
Oct. 1975	Telecom 75	Paris → Pleumeur Bodou	Geneva (Switzerland) (mobile station SAMSON)	Symphonie	Differential demodulators (see Section IV-C-1)
		Picturephone			
Nov. 1976	19th UNESCO General Conference	Paris Unesco → Pleumeur Bodou Headquarters	Nairobi (Kenya) (mobile station SAMSON)	Symphonie	
		Picturephone or 30 PCM voice channels + 6 telegraphic channels			
Since 1978	In normal Operation	Pleumeur-Bodou	La Rivière des Pluies (I. of La Réunion) (satellite earth station)	Symphonie or Intelsat 4A	
		2×30 PCM voice channels			
July 1980	Test link	Pleumeur-Bodou	La Rivière des Pluies 240 PCM voice channels	Intelsat 4A	Digital modulator (see Section II-A)
Sept. 1980	In normal operation	Bercenay-en-Othe (France) (satellite earth station)	La Rivière des Pluies 240 PCM voice channels	Intelsat 4A	IF transmit and receive filters (see Sections III-B and IV-B) Coherent demodulator (see Section IV-C-2)

very-narrow-bandwidth minimum phase HF filters for carrier recovery circuits. In addition, SAW techniques can supply 2- or 4-PSK all-digital modulators with fixed frequency oscillators, and up and down frequency converters with VCO's. This results from their adaptiveness to the requisite frequency ranges and from their favorable stability as well as electronic tuning properties.

Finally, SAW devices are remarkably well suited to the needs of satellite digital communication systems. They call for an advanced technology—conventional, but sometimes submicrometer microphotolitography—and in this regard define a new generation of equipment with advanced components.

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