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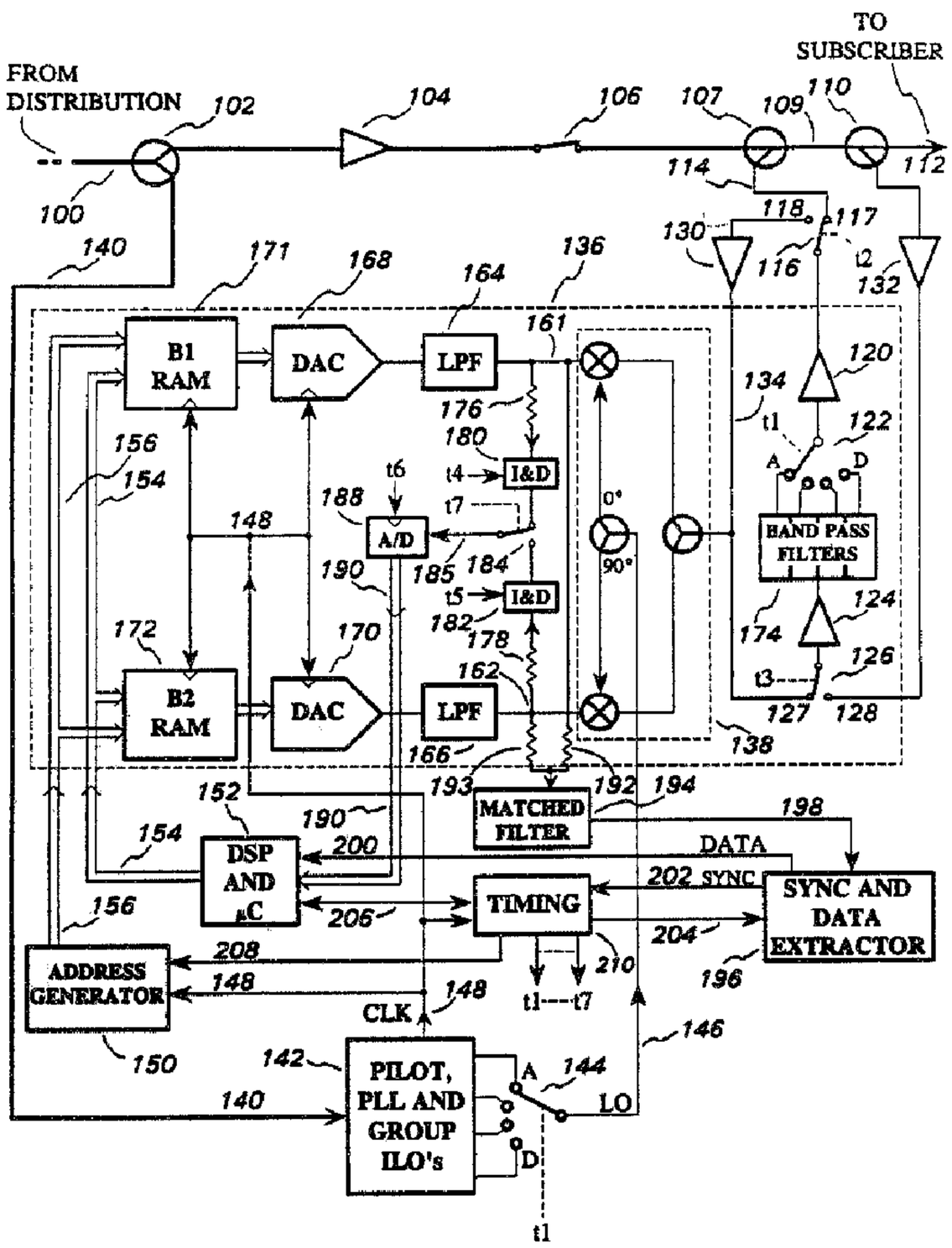

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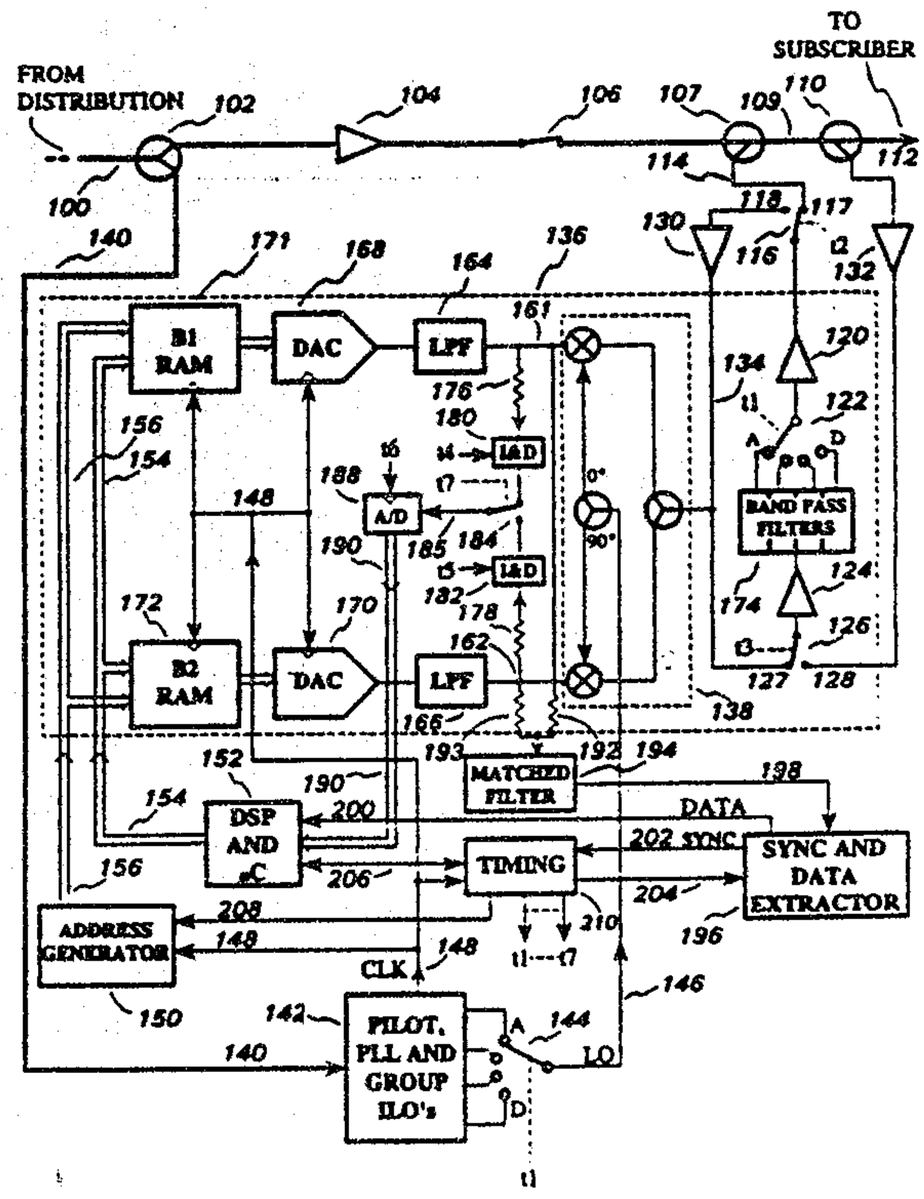
(54) Title: MULTICHANNEL TELEVISION SIGNAL SCRAMBLING AND DESCRAMBLING SYSTEM AND METHOD

(57) Abstract

A multichannel television access control system is disclosed wherein processing of the broadbande signal is performed by gated coherent RF injection on a plurality of channels, thereby effecting selective descrambling of an arbitrary subset of channels in a channel group. The system provides for a closed loop injection signal phasor control so as to maintain substantially constant relationship between the television signals and the injected signals. The combined signals are supplied to subscribers enabling the use of their cable ready equipment. The broadband signal is coupled to a broadband amplifier (104) via a power splitter (102). The injected signals are combined through line (114) with the amplified broadband signal by directional coupler (108), resulting in the composite signal on line (109). The broadband signal on line (109) is fed to the subscriber port (112) through directional tap (110). The injected signals forming together the composite signal injected at line (114) are generated by a digital broadband generator (136) feeding RF switch (116). During the time in which injection is required, RF switch (116) is at position (117) as shown in the figure.



MULTICHANNEL DIGITAL SIGNAL GENERATION METHOD AND APPARATUS

FIELD OF THE INVENTION

The present invention relates generally to television transmission and receiving systems and more particularly to CATV or over-the-air multichannel systems wherein a plurality of standard television signals are scrambled by suppression of normal synchronizing pulses or otherwise modified during the active video time and subsequently selectively descrambled at authorized subscriber locations in a manner which simultaneously renders a plurality of authorized channels to be provided unscrambled to the subscriber's television set.

BACKGROUND OF THE INVENTION

In subscription television, or pay television, programming signals of premium channels are typically transmitted in a scrambled form so that unauthorized viewers, that is, viewers who do not pay to receive the specific programming, are not able to view the transmitted signal on their television receivers. Authorized viewers are provided with means to descramble or decode the scrambled television signal and usually such descrambling means are capable of being selectively enabled to descramble specific channels. This enablement can be done remotely by the service provider or the CATV operator, by selective addressing of individual descrambling means associated with subscribers who paid to receive the scrambled programming.

A technique commonly used for television scrambling is that of sync-suppression wherein the RF level of the horizontal and vertical synchronizing pulses is suppressed to a level below that of the video so that the standard television receiver is unable to establish regular synchrony and instead erratically locks on RF peaks in the active video

thereby creating unviewable picture on the television set. In addition, the ability of the television receiver to use the color reference burst associated with the horizontal synchronizing pulse is severely degraded thereby causing distorted color reproduction or disabling the color circuitry of the television set. Exemplary prior art sync suppression systems are disclosed in U.S. Pat. Nos. 3,184,537 to Court et al; 3,478,166 to Reiter et al; 3,530,232 to Reiter et al and 4,222,068 to Thompson. In these systems, the horizontal synchronizing levels are suppressed or reduced to gray level and an additional keying control signal is normally transmitted together with the scrambled television signal for controlling reconstruction or regeneration of the proper sync levels at the authorized descrambler, thereby providing unscrambled normal television signal at the receiver. For example, in U.S. Pat. No. 3,184,537 the audio subcarrier is amplitude modulated with a suitable sync insertion control signal. In U.S. Pat. No. 4,222,068 horizontal sync reconstruction at the authorized descrambler is effected by transmitting normal sync signals during the Vertical Blanking Interval ("VBI") of the television signal for enabling a timing circuit at the receiver to lock on the horizontal components thereof. The timing circuit is then used to gate the Horizontal Blanking Interval ("HBI") thereby facilitating the restoration of the horizontal synchronizing pulses in the composite baseband video signal.

Sync suppression systems normally operate in one of two ways: The first method, as in U.S. Pat. No. 3,184,537, is based on attenuating the modulated RF television signal during the HBI (and in some cases during the VBI as well) by a known fixed amount such as 6 Db. At the authorized descrambler, the signal is descrambled by attenuating the RF received signal during the active video portion by the same known amount. This method is known as "RF sync suppression" since the scrambling and descrambling processes occur at RF stages after the modulation process at the transmitter and before the demodulation process at the receiver respectively. The second sync suppression method is known as "Baseband Sync Suppression" wherein a known gated offset level is added to the baseband video signal during the HBI, thereby suppressing the horizontal synchronizing signal. At the authorized descrambler, an equal offset level is subtracted from the demodulated video signal during the HBI, thereby restoring the

horizontal synchronizing signal levels to their normal values. An example of Baseband sync suppression system is disclosed in U.S. Pat. No. 4,222,068.

Both prior art RF and baseband sync suppression descrambling systems described above are typically accomplished by incorporating the timing and descrambling circuitry with a television tuner and demodulator, often termed a converter, that can cover the entire CATV band or over-the-air channels used by the subscription television service provider. In RF sync suppression systems, descrambling is effected by gated attenuation at the IF stage of the converter or at its fixed RF output channel frequency such as TV channel 3 or 4. In baseband sync suppression systems, descrambling is effected at the demodulated video stage of the converter and the resultant normal video signal is remodulated by means of a fixed frequency television modulator operating on channel 3 or 4. Thus, in both examples, the subscriber is equipped with a converter-descrambler device capable of tuning and descrambling a single channel at a time and converting it to channel 3 or 4 (output channel), whichever is unused by a local broadcast station. The subscriber's television set may permanently remain tuned to the converter-descrambler output channel while channel selection is done by means of the converter-descrambler tuner.

In a subscription television service, a plurality of scrambled channels may be transmitted simultaneously on several channels. However, as explained above, prior art converter-descramblers can only process one channel at a time. This means that subscriber households who are authorized to receive several premium scrambled channels, cannot receive these services descrambled simultaneously. So for example, if dad wishes to watch a "pay-per-view" event such as a championship fight, while the kids watch the Disney Channel on the second television set, the household would require two converter-descramblers. Furthermore, if mom wanted to record a movie on HBO at the same time, a third converter-descrambler must be employed in conjunction with the Video Cassette Recorder ("VCR"). This is clearly one of the disadvantages of prior art single channel descrambling techniques.

Most television households in America are equipped with "Cable ready" television sets or VCR's. The term "Cable ready equipment" should be understood as consumer television equipment with expanded frequency tuning capability that allows a

subscriber to tune not only the off-air broadcast channel frequencies but also all CATV channels. A significant portion of the cable ready equipment sold in America contains additional features such as Multichannel Television Sound ("MTS") allowing reception of stereo TV programming or second audio programming ("SAP"), remote control to control sound level or sound muting as well as channel tuning. Still another feature of some cable ready television sets is the Picture-In-Picture ("PIP") display capability allowing the subscriber to simultaneously watch one program while observing additional video program(s) in a picture inset displayed within the main picture. When this cable ready equipment is connected directly to the cable drop entering the household, the subscriber is usually able to receive and record all non-scrambled channels without the aid of a converter and thus use his equipment's special features described above to receive such non-scrambled signals. The problem occurs when the CATV system employs signal scrambling as means of controlling access to the premium channels.

Normally in these circumstances, being fed by the converter-descrambler's output channel, the television set is permanently tuned to channel 3 (or 4) while the converter-descrambler is tuned to the desired channel. Thus, the remote control purchased with the television set is unused, and an additional remote control unit must be provided for the converter-descrambler. Moreover, many prior art baseband converter-descramblers cannot pass the composite MTS audio program material and thus the stereo signal and the SAP are lost despite the fact the television set may be capable of receiving MTS. Recent attempts to address this problem typically involve a stereo bypass operating mode as disclosed in U.S. Pat. No. 4,630,113 to Long, which unfortunately disables the remote volume control of the converter, or alternatively by dematrixing and reprocessing the stereo signal at the converter in order to effect volume control with resultant degradation of stereo separation and compander performance. Even prior art RF sync suppression descramblers tend to introduce MTS audio performance degradations since they pass the audio subcarrier with additional amplitude modulation at the horizontal line rate thereby causing AM to FM effects at the television receiver. These audio degrading effects are discussed in an article by J. O. Farmer entitled "Operational Characteristics of Modern Set-Top Terminals" published in the *Technical Papers of the 33rd Annual Convention of the NCTA* in Las Vegas, June 3-6, 1984 and

in the *IEEE Transactions on Consumer Electronics*, **Vol. CE-30**, pages 489-502, August 1984. Finally, the PIP feature of new television sets requires that the TV set be able to receive multiple channels simultaneously, a requirement that obviously cannot be met by current prior art converter-descramblers. Clearly, all these problems constitute yet another set of significant disadvantages of the prior art single channel descrambling techniques.

Collectively, the foregoing deficiencies discussed above are known in the industry as the "CATV consumer interface" problem. There is a growing body of evidence that these consumer interface problems present severe hardships on consumers and CATV operators alike. Specific evidence and indication of the public interest in this regard is expressed by legislation cited as the "Cable Ready Equipment Act of 1991" recently proposed by Senator Patrick Leahy in U.S. Senate Bill S. 2063, published in the Congressional Record - Senate, pages S18377 - S18380, November 26, 1991. Leahy's bill seeks to encourage solutions to the CATV consumer interface problems described above.

It may be easily understood that the most desirable solution to the problems discussed above is to provide the subscriber with a cable drop which carries all the channels to which he/she is entitled in unscrambled form ("in the clear"). The subscriber can then connect by proper broadband RF signal splitting means multiple cable ready television sets and VCR's and freely enjoy all of the consumer features that were purchased with that equipment. Since the CATV system may be carrying other channels not purchased by the subscriber, an effective means is required to limit a subscriber's access only to those channels and special events for which subscription fees have been paid.

If one attempts to employ prior art single channel descrambling techniques in order to provide each subscriber with simultaneous multichannel descrambling, one would have to provide each subscriber with a multiplicity of prior art descrambling means wherein each descrambler is tuned and dedicated to one scrambled channel. The outputs of such descramblers must then be assigned distinct channel frequencies and must be combined by means of a diplexer with all non-scrambled channels to form a broadband multichannel clear signal service entering the subscriber home. Clearly, the

cost per subscriber for such an arrangement could reach the prohibitive value of the cost of one descrambler times the number of processed channels.

There thus exists the need for an effective low cost system for simultaneously descrambling an arbitrary subset of a plurality of scrambled channels, said arbitrary channel subset being the channels to which the subscriber has subscribed, and supply them together with the basic non-scrambled channels into the home in such a manner that all authorized channels are supplied 'in the clear' and all scrambled channels not purchased by the subscriber are passed through into the home either in their original scrambled form, or preferably with additional security added to further deny any unauthorized access by "piratical" devices.

SUMMARY OF THE INVENTION

The present invention is directed to subscription television system employing video scrambling on a plurality of channels. A broadband multichannel simultaneous processing system is disclosed which performs descrambling and other access control functions on a plurality of addressable selected channels in accordance with the subscriber authorization to receive each channel.

It is an object of the present invention to provide a simultaneous multichannel television descrambling system of the type described, enabling the cable ready and other consumer features contained in the subscriber's television sets and VCR's to be fully utilized by the subscriber.

It is a further object of the present invention to provide a scrambling and descrambling system of the type described which introduces virtually no artifacts or distortion in either the audio or video signals on descrambled and non-scrambled channels, wherein video quality is not compromised and functions such as MTS stereo are not degraded or affected.

It is still another object of the invention to provide a subscription television access control system that offers CATV system operators the ability to migrate currently used prior art systems to a new encryption based multichannel video scrambling method with an improved head end originated security scheme providing an enhanced security. Such system shall allow migration to the new encryption based security system on any channel for which all authorized subscribers are served by an installed access control device of the present invention.

To that end and other objectives, a method for processing a broadband signal comprised of a plurality of amplitude modulated television signals comprising the steps of generating a plurality of radio frequency (RF) signals in phase synchrony with at least a portion of said television signals, each of said plurality of of generated RF signals having phasor values; combining said broadband signal with said generated RF signals, and adjusting said phasor values of each of said plurality of generated RF signals in accordance with corresponding phasor values of said television signals so as to maintain a substantially constant relationship therebetween.

Accordingly, in one aspect, the present invention provides a method of generating a broadband signal comprising up to $2N+1$ RF signals where N is an integer value, each of the RF signals having a frequency that is related to a center frequency by an integral multiple of a positive or negative increment of a fundamental frequency, each of the RF signals having an amplitude and phase characterized by a phasor value, said method comprising the steps of: generating a first and second baseband signal each comprising a linear combination of up to N sinusoids, each of the sinusoids having a frequency that is an integral multiple of the fundamental frequency; and modulating in quadrature the first and second baseband signals on a third signal having a frequency equal to the center frequency.

In a further aspect, the present invention provides an apparatus for generating a broadband signal comprising up to $2N+1$ RF signals each having a frequency separated from a center frequency by an integral multiple of a positive or negative increment of a fundamental frequency, said apparatus comprising: memory means for storing a sequence of digital samples representing first and second baseband signals, each of the baseband signals comprising a linear combination of up to N sinusoids, each of the sinusoids having a frequency that is an integral multiple of the fundamental frequency; address generator means, coupled to an address port of said memory means, for generating addresses accessing the sequence of digital samples from said memory means; clock signal generator means, coupled to said address generator means, for varying the address values generated by said address generator means, said clock signal generator means having a frequency equal to an integral multiple of the fundamental frequency; digital to analog conversion means, coupled to said memory means, for converting the accessed digital samples to the first and second baseband signals; signal generation means for generating a local signal having a frequency equal to the center frequency; and modulator means, coupled to the first and second baseband signals and the local signal, for modulating the first and second baseband signals on the local signal to produce the broadband signal.

In a further aspect, the present invention a method of generating a broadband signal comprising up to $2N+1$ RF signals where N is an integer, each of the RF signals having a frequency that is related to a center frequency by a frequency difference that is a positive or negative integral multiple of a fundamental frequency, each of the RF signals having an amplitude and phase characterized by a phasor value, said method comprising the steps of: generating a first and second baseband signal each comprising a linear combination of up to N sinusoids, each of the sinusoids having a frequency that is an integral multiple of the fundamental frequency; and modulating in quadrature the first and second baseband signals on a local signal [146] having a frequency equal to the center frequency.

In another aspect, the present invention provides an apparatus for generating a broadband signal comprising up to $2N+1$ RF signals where N is an integer, each of the RF signals having a frequency separated from a center frequency by a frequency difference that is a positive or negative integral multiple of a fundamental frequency, said apparatus comprising: means for generating digital samples representing first and second baseband signals, each of the baseband signals comprising a linear combination of up to N sinusoids, each of the sinusoids having a frequency that is an integral multiple of the fundamental frequency; means, coupled to said means for generating digital samples, for converting the digital samples to the first and second baseband signals; means for generating a clock signal coupled to said means for generating digital samples and to said means for converting; means for generating a local signal having a frequency equal to the center frequency; and means, coupled to the first and second baseband signals and the local signal, for modulating in quadrature the first and second baseband signals on the local signal to produce the broadband signal.

BRIEF DESCRIPTION OF DRAWINGS

To the accomplishment of the above and such further objects as may hereinafter appear, the present invention relates to a television access control system substantially as defined in the appended claims, and as described in the following detailed specification as considered in conjunction with the accompanying drawings, in which:

Figure 1a is a baseband sync suppression waveform;

Figure 1b is the coherent RF injected signal for descrambling the signal of Figure 1a in accordance with the present invention;

Figure 2a is the spectrum of the television channels entering the multichannel descrambler of the present invention;

Figure 2b is the spectrum of the injected signals generated by the broadband generator of the present invention;

Figure 2c is the descrambled broadband spectrum supplied to the subscriber from the multichannel descrambler of the present invention;

Figure 3 is the first preferred embodiment of the present invention;

Figure 4 is a plot of two baseband quadrature components required for VSB spectral shaping of the injected signals used for multichannel descrambling in accordance with the present invention;

Figure 5 is the computational digital filter used to generate the quadrature components of Figure 4;

Figure 6a describes the broadband coverage by channel group time sharing of the injected signals of the present invention;

Figure 6b is the frequency spectrum partitioning by channel block grouping;

Figure 6c is the frequency spectrum partitioning by frequency interlaced grouping;

Figure 7 is the timing diagram of coherent injection groups and their respective RAM address schedule;

Figure 8 is a phasor diagram describing the received signals and the injected signals;

Figure 9 is an active calibration and analysis timing diagram in the preferred embodiment of the present invention;

Figure 10 is another active calibration and analysis timing diagram including RAM address space trajectory;

Figure 11 is an analysis and measurement timing diagram describing the phasor tracking mechanism;

Figure 12 is an overall video timing diagram describing pertinent events and processes utilized in the present invention;

Figure 13 is an example of a CATV head end configuration desired for the practice of the present invention;

Figure 14 is a further detail of the comb generator system utilized at the headend;

Figure 15a is a baseband representation of an RF sync suppression waveform;

Figure 15b is the coherent RF injected signal for descrambling the signal of Figure 15a in accordance with the present invention;

Figure 16 is a digital broadband generator for the second preferred embodiment of the present invention;

Figure 17 is a second preferred embodiment of the present invention directed towards use in multiple dwelling units;

Figure 18 is a timing and memory diagram for an enhanced security video scrambling using the injection methods of the present invention.

DETAILED DESCRIPTION OF PREFERRED EMBODIMENTS

It should be appreciated that since the object of the present invention is to provide all channels to the subscribers in a broadband form and that only certain channels must be descrambled, and hence modified, the approach taken according to the present invention is to avoid any single channel filtering or tuning techniques but rather focus on techniques that involve broadband signal addition in such a manner as to modify (descramble) only selected channels. From the outset, it is instructive first to observe that both baseband sync suppression and RF sync suppression scrambling constitute a linear modifying process in the radio frequency domain. This linear process is active only during the blanking intervals in which suppression occurs. This means that for each of these two scrambling process, there exists an additive RF signal pulse of the appropriate onset time, duration, amplitude envelope, frequency and phase, such that when added to the scrambled RF signal, results in an RF television signal with normal synchronizing signals, and hence unscrambled.

Figure 1a depicts the baseband video signal during the HBI in a baseband sync suppression system. As can be seen, the normal sync signal **10** is suppressed by adding at the scrambler a baseband offset signal of magnitude **20** during the HBI which results in suppressed sync signal **12**, while the active video signal **14** is unmodified. Furthermore, as can be seen in Figure 1a, the color reference burst signal **16** is unmodified but merely superimposed on an offset level of magnitude **20** in the figure. To those skilled in the art of television modulation, it is well known that the sync level **10** (-40 IRE units) corresponds to peak RF amplitude while video white level **22** (100 IRE level) corresponds to an RF amplitude that is 12.5% of peak RF amplitude. Thus, in this inverted modulation scale of the RF domain, the HBI gated offset level **20** at baseband is equivalent to the subtraction of an RF pulse shown in Figure 1b from an otherwise non-scrambled RF television signal. The frequency and phase of this RF pulse is equal to those of the picture carrier of the television signal. Hence, in order to restore a baseband sync suppression signal corresponding to **12** to it's unscrambled mode, one can add a coherent RF pulse train with the horizontal synchronizing repetition rate, coinciding in time with the HBI as shown in Figure 1b, with the appropriate amplitude **30** and in phase with the picture carrier so as to obtain an unscrambled signal, the

baseband representation of which corresponds to 10. During the active video time 34, essentially no RF signal must be present.

Of course such addition at RF, hereinafter termed "Coherent RF Injection", must be effected with appreciable amplitude, phase and timing precision so as to obtain essentially an unscrambled signal. For example, a common baseband sync suppression system used in the American CATV market is essentially in accordance with U.S. Pat. No. 4,222,068 wherein the baseband sync level shift **20** is set to 70 IRE units. Thus, in this case, using the standard television modulation scale, it can be shown that the in-phase coherent RF injection level **30** required to effect descrambling must have an amplitude corresponding to 44.75% of the peak RF level of the non-scrambled portion of the signal. Moderate errors in the level of such injected RF pulse may not affect the proper synchronizing of the television receiver for satisfactory viewing of unscrambled programs, but may cause errors in the black reference level of the television set. This is because most sets establish their black video reference level by sampling the 0 IRE level **24** during the color burst **16** period. However, generally accepted video specifications embodied in the well known RS-170A standard allow for fixed errors up to $\pm$ 2 IRE in black level. These error levels are permissible and are undetected for all practical purposes.

Similarly, the injection phase of the RF pulse must be within approximately $\pm 6°$ of the picture carrier phase. Because the injected RF amplitude is nearly equal to that of the suppressed sync amplitude, these injection phase errors would result in total false phase modulation of the picture carrier by no more than $\pm 3°$ during the HBI. This requirement limiting the phase error to $\pm 3°$ stems from two different considerations: The first is related to the fact that good broadcast television engineering practice requires that Incidental Phase Modulation ("ICPM") shall not exceed 3° so as not to cause perceptible intercarrier audio buzz or MCS stereo distortion. The second reason for this accuracy requirement stems from the fact that this phase error only exists during the HBI, within which the color reference burst signal **16** is received, and not during the active video time **14**. The RF phase shift during the HBI causes an equal baseband phase shift in the detected color reference signal, thereby causing color errors in the detected chrominance signal during the active video. Subjective tests evaluating the permissible color reference

phase errors in NTSC television signals were conducted by Bernard D. Loughlin of the Hazeltine Corporation and reported in a chapter entitled "Nonlinear Amplitude Relations and Gamma Correction" on pages 63-91 in a book edited by Ted Rzeszewski entitled *Color Television* published by the IEEE, 1984. Figure 11-27 in this chapter shows that phase errors of $\pm 3°$ are not perceptible by at least 80% of observers.

Based on the permissible amplitude and phase error magnitudes discussed above, it can be shown that the required accuracy of coherently injected RF signals for descrambling can be characterized by an ideal errorless injection signal accompanied by an injection RF phasor error of magnitude not exceeding -20 Db as compared to the desired injected RF signal. As will be subsequently appreciated, this permissible error level facilitates a low cost embodiment of the present invention for simultaneous injection of a plurality of channels.

Reference is now made to Figure 1a in which the onset and termination of sync suppression for every horizontal video line at baseband is shown as level transitions **26** and **27** respectively. These baseband transitions at the scrambler take place over a time duration not exceeding two hundred nanoseconds, consistent with a standard television video bandwidth of 4.2 Mhz. Upon descrambling of such signals, it is preferable to match these transition times when the sync offset level is removed at the descrambler in accordance with the present invention by coherent injection of an RF pulse shown in Figure 1b. Here, the onset and termination periods of the injected RF pulse are designated by **32** and **33** respectively. Thus, their durations would preferably each be no longer than two hundred nanoseconds.

If the pulse shape of the injected RF envelope of Figure 1b is attempted by means of an amplitude pulse modulator with video bandwidth, the fast rise-times **32** and **33** would generate double sideband spectral broadening of up to 4.2 Mhz above and below the picture carrier frequency. The upper sideband content of this injected signal will be contained within the desired normal television bandwidth as transmitted in Vestigial Sideband ("VSB") Modulation. However, since the injected signal is assumed to be combined with the received signal in a broadband combiner, the lower sideband of the injected signal may interfere with a lower adjacent television channel and in particular with it's audio subcarrier located only 1.5 Mhz below the picture carrier of the

descrambled channel. If, instead, longer transition times are assigned for 32 and 33 so as to limit the spectral broadening of the amplitude modulated pulse to less than 1.5 MHz, the descrambled video signal will contain front and back porch transients with durations in the microsecond range, which may invade the leading edge of the horizontal synchronizing signal 12 or the active video time 14 of the descrambled signal, thereby delivering a degraded video signal to the subscriber which may cause false horizontal sync or unstable video clamping action by the television set.

These two conflicting requirements in the frequency domain and in the time domain as discussed above, can be resolved by turning to the very method which allow fast video transitions in television transmission without lower sideband spectral expansion, namely the use of VSB modulation techniques for the transitions 32 and 33 prior to the RF combining with the scrambled signal. This will allow the television receiver to process the upper sideband containing up to 4.2 MHz wide energy associated with fast baseband transitions while limiting the lower sideband expansion below 1.5 MHz and thus preventing the associated interference to the lower adjacent channel. An elaboration on the digital embodiments which facilitate such VSB spectral shaping of the injected signal will be found in subsequent sections below.

The coherent gated RF injection described above can be useful for purposes other than descrambling. Provisions are made in the present invention for implementing signal denial techniques for non authorized subscribers on a channel by channel basis. In contrast to the descrambling case, in these signal denial cases, the coherent gated RF injection of Figure 1b can be effected in opposite phase to that of the picture carrier of the television signal, thereby further suppressing or nulling the synchronizing signals and optionally, with sufficient injection level, even reversing the resultant RF phase of the received television signal during the HBI. This method results in enhanced security for unauthorized subscribers since it denies "pirate" decoders the ability to reconstruct the sync signals and further causes phase discontinuity in the intercarrier audio detector of the television set, thereby introducing disturbing audio buzz and further audio noise masking in some television sets. It should be understood that coherent RF injection for denial of the type described above must also be gated during the HBI using VSB

transition modulation so as to prevent interference to the lower adjacent channel that may otherwise be clear or authorized for descrambling.

Thus far, processing of the television signals at RF by coherent injection was discussed in a context of a single channel. Clearly, the main reason to use RF injection techniques is to offer simultaneous processing of this novel type for a plurality of channels. This means that coherent injection should take place at each frequency for which either descrambling or further sync denial is required.

Figure 2a depicts a portion of the broadband spectrum of the incoming signal carried on the CATV system entering the multichannel descrambler of the present invention. Each television channel contains a picture carrier **40** and an audio subcarrier **42** separated by 4.5 MHz from the picture carrier **40**. The channel spacing **48** is 6 MHz. As can be seen, the spectra of the television signals appears asymmetric about its picture carrier frequency as it is upper sideband VSB modulated. In Figure 2a one can observe two types of channels being transmitted down the CATV distribution. The first constitutes all the non-scrambled channels **46** for which all subscribers are authorized and thus we assume no processing is required. The second type of channels carried is the sync suppressed scrambled channels **44**, designated with the letter S. In this example the assumption is made that the subscriber has purchased subscriptions for a portion of the premium scrambled channels **44**, said channel portion designated here by **44'**. Consequently, the required RF injection spectrum corresponding to this subscription configuration is depicted schematically in Figure 2b. Each signal in Figure 2b is assumed to be generated within the multichannel descrambler of the present invention and combined in phase lock with its respective transmitted counterpart picture carrier of the same frequency in Figure 2a. The broadband composite signal of Figure 2b is coherently injected and thus linearly combined with the broadband incoming signal of Figure 2a to form the composite broadband signal depicted in Figure 2c that is subsequently provided to the subscriber for his/her viewing pleasure. The injected signals **50,52** consists of VSB injection signals with time domain profiles of the type depicted in Figure 1b. The injected RF signals **50** are injected in phase with respect to their corresponding incoming scrambled signals **44'**, thereby effecting simultaneous descrambling and resulting in clear channels **60** of Figure 2c being provided to the

subscriber. In contrast, the injected RF signals **52** are injected out of phase, thereby causing sync null or other signal denial effects such as sync phase reversal. The resultant television signals **62** are not viewable or otherwise useful to the unauthorized subscriber. However, because of the broadband characteristics of this combining system, they are fed to the subscriber along with all other clear signals **60**.

Because the injected signals **50** and **52** all have identical temporal profile as that of Figure 1b, their simultaneous generation as a group can be made much simpler if they are all required at the same time, that is if the HBI of all television signals in the channel group of Figure 2a coincides in time repeatedly. This relative timing coincidence condition among this group of video channels is well known to those skilled in the art of video engineering as "frame synchronous" video sources. This condition can be accomplished at the CATV headend before each channel is modulated by means of video frame synchronizers providing video outputs genlocked to a master video synchronizing source.

It will become apparent in following discussions regarding preferred embodiments of the present invention that another appreciable subscriber injection generator simplification can be realized in CATV systems that employ picture carrier frequency assignments and control in accordance with an Incrementally Related Carrier ("IRC") or Harmonically Related Carrier ("HRC") channel plan, at least within that channel portion of the band that is being processed in accordance with the present invention. This simplification is related to the fact that under such conditions, the composite RF signal of Figure 2b can be derived from a periodic signal with fundamental periodicity equal to the Incremental frequency **48** separating any two nearest channels within the group. This RF synchrony condition can be effected at the CATV headend by phase locking all modulators in the channel group to an appropriate IRC or HRC comb signal as disclosed for example by the present inventor in U.S. Patent No. 5,125,100 and in the cited references therein, or in U.S. Patent No. 4,099,123 to Finlay et al.

A first preferred embodiment of the present invention is depicted in Figure 3. The assumption is made that the headend video frame synchronization and IRC or HRC carrier phase locking is effected as discussed above. The broadband incoming signal containing all transmitted channels arrives on line **100**. A portion of the power of the

broadband signal is coupled to a broadband amplifier **104** via a power splitter **102**. When the subscriber is authorized, RF switch **106** is closed and the amplified broadband signal corresponding to Figure 2a feeds the directional coupler **108**. The injected signals corresponding to Figure 2b, each having a temporal profile in accordance with Figure 1b, are combined through line **114** with the amplified broadband signal by directional coupler **108**, resulting in the composite signal of Figure 2c on line **109**. The broadband signal on line **109** is fed to the subscriber port **112** through directional tap **110**. The injected signals **50** and **52** of Figure 2b forming together the composite signal injected at line **114** are generated by a digital broadband generator **136** feeding RF switch **116**. During the time in which injection is required, RF switch **116** is at position **117** as shown in the figure.

The Digital broadband generator **136** is capable of generating signals **50** and **52** situated on arbitrary subset of frequency points on a 6 MHz grid centered about a center local oscillator frequency fed on line **146**. These signals are generated by means of quadrature modulator **138** being fed with appropriate baseband signals on lines **161** and **162** and having a one sided bandwidth not exceeding the cutoff frequency of low pass filters **164** and **166**. The signals thus generated are amplified by broadband amplifier **124** and fed through the appropriate SAW band pass filter in filter bank **174**, to broadband amplifier **120** and to RF switch **116**. The baseband signals feeding the two quadrature inputs **161** and **162** are functions of time designated hereinafter as $B_1(t)$ and $B_2(t)$ and are digitally generated by Digital to Analog Converters ("DAC") **168** and **170** respectively, based on the contents of their respective RAM's **171** and **172**. Low pass filters **164,166** provide antialiasing filtering as well as out of band harmonic rejection. The clock frequency driving DACs **168,170** and RAMs **171,172** is fed at line **148**. This clock signal also advances the RAM address generator **150**, causing the appropriate RAM data contents to be loaded sequentially into DACs **168** and **170** respectively. The clock frequency is set so that it is a sufficiently large integral multiple of the fundamental frequency increment **48** between two adjacent channels. For a 6 MHz channel spacing one might select the clock frequency to be 72 MHz, thereby allowing digital generation of baseband signals approaching the Nyquist rate of 36 MHz. With practical low pass

filter designs, one can easily generate injection signals for a 9 channel group as shown in Figure 2.

Referring again to Figure 1b, it should be noted that during the time interval **30**, the amplitude and phase of the injected signals are constant, thereby requiring a CW sinusoidal signal to be injected at each carrier frequency for which additive processing is required. The assumption is made that the local oscillator feeding the quadrature modulator **138** on line **146** is phase locked to a channel group center carrier arriving on line **140** by an Injection Locked Oscillator ("ILO") or Phase Locked Loop ("PLL") oscillator contained in subsystem **142**. A further assumption is made that the clock signal on line **148** (72 MHz in this example) is similarly phase locked to a pilot signal arriving on line **140** which is an integral multiple of the fundamental incremental frequency **48** (6 MHz in this example). Due to the linearity of the quadrature modulation process in **138**, the required signals $B_1(t)$ and $B_2(t)$ must be each a linear combination of sinusoids with frequencies that are integral multiples of 6 MHz. Hence, during that period of time, $B_1(t)$ and $B_2(t)$ are periodic waveforms with a 6 MHz periodicity, thereby allowing a discrete time representation of each signal with M samples (wherein in the present example $M = 72\ \text{MHz}/6\ \text{MHz} = 12$). Therefore, the RAM address generator **150** need only repetitively scan through 12 RAM addresses during period 30 in order to generate an arbitrary CW frequency grid at **134**. The specific amplitude and phase of the generated injection carriers will depend on the values of the 12 scanned samples in the B1 RAM and the 12 scanned samples in the B2 RAM.

At the RF domain, each generated carrier output on line **134** can be resolved into two quadrature components I and Q. The index k shall be used here to designate the baseband order of the 6 MHz harmonic corresponding to the separation of the specific carrier in question from the center frequency carrier on which the group local oscillator is locked. Thus, for any given k (up to a value of N), there are two sideband components that can be generated, corresponding to two distinct channels. Given their desired quadrature components I_k^+, Q_k^+, I_k^- and Q_k^- represented by a 4 entry column vector, the baseband signals $B_1(t)$ and $B_2(t)$ (represented by a 2 entry column vector for each sample time t) can be shown to be given by the following vector equation:

$$\begin{bmatrix} B_1(t) \\ B_2(t) \end{bmatrix} =$$

$$= \sum_{k=0}^{k=N} \begin{bmatrix} \cos(2\pi k\frac{t}{M}) & \sin(2\pi k\frac{t}{M}) & 0 & 0 \\ 0 & 0 & \cos(2\pi k\frac{t}{M}) & \sin(2\pi k\frac{t}{M}) \end{bmatrix} \begin{bmatrix} 0 & 1 & 0 & 1 \\ 1 & 0 & -1 & 0 \\ 1 & 0 & 1 & 0 \\ 0 & -1 & 0 & 1 \end{bmatrix} \begin{bmatrix} I_k^- \\ Q_k^- \\ I_k^+ \\ Q_k^+ \end{bmatrix} \quad (1)$$

With an arbitrary RF phase reference coordinate system in Equation (1), I_k^+ and Q_k^+ correspond respectively to the in-phase and quadrature components of the upper sideband RF injection signals of frequency index k, while I_k^- and Q_k^- correspond to those of the lower sideband of frequency index k. The parameter t is the discrete time variable assuming integer values of 1 through M. In the case $k=0$, both sidebands degenerate into two halves of a single center carrier, for which Equation (1) reduces to an equation for 2 entry vectors and a 2x2 matrix.

By proper calculations essentially in accordance with Equation (1), based on the required generated values of I_k^+,Q_k^+,I_k^- and Q_k^-, the Digital Signal Processor and Micro Controller ("DSP/uC") **152** evaluates the required samples $B_1(t)$ and $B_2(t)$, and stores them in RAMs **171,172** via data bus **154**. As indicated above, the injected signals corresponding to the column vector in the extreme right hand side of Equation (1) must be fed in proper phase and amplitude so that when observed on line **109** they match the phasor values of the incoming signals also observed on the same line **109**. To this end, the present invention provides for phasor measurement and analysis means that preferably may be comprised of essentially the same apparatus means that effect the coherent injection, namely, digital broadband generator **136**, whereby these two functions are time shared in accordance with a predetermined video timing schedule. It is possible to time share and perform measurements during time periods in which no descrambling or RF injection is required. These measurement or phasor analysis periods would preferably be within the VBI, during time intervals in which no active video

information is present, therefore providing a fixed received reference for the picture carrier level and phase.

Turning now to these phasor analysis time periods for which no injection is required, an RF switch **116** is set in position **118** and RF switch **126** is turned to position **128**. In this way, a sample of the signal to be analyzed on line **109** is fed through a directional tap **110** then amplified by the broadband amplifiers **132, 124** feeding the appropriate filter in filter bank **174**. This filtered signal is then further amplified by the broadband amplifiers **120,130** then fed to the quadrature modulator **138** at line **134.** It is parenthetically noted that the assumption made here is that the quadrature modulator **138** is reciprocal, that is, that it is constructed from passive devices allowing its internal power combiner to act as a power splitter and that both it's balanced mixers are bidirectional. Under these assumptions, device **138** may also be referred to as a quadrature demodulator, for which the quadrature outputs are on lines **161,162.** An example of a reciprocal quadrature modulator **138** that may be employed in accordance with the teaching of the present invention is Model QMC-170 manufactured by the Mini-Circuits corporation of Brooklyn, New York. It can be shown that by providing a particular sinusoidal signal pair of a given frequency *k* at the output of the low pass filters **164** and **166** and feeding them to the mixers of the quadrature demodulator **138** at lines **161**, and **162** respectively, double mixing will occur, so that DC components related to the secondary mixing at baseband will result at lines **161**, and **162.** This operation is essentially that of a double conversion quadrature receiver where the first local oscillator is supplied at line **146** while the second "local oscillators" are supplied in quadrature by pure sinusoidal signals digitally generated by the DACs **168** and **170** through lines **161** and **162**. The DC components on these same lines constitute the detected signals coupled through high impedances **176** and **178** to a gated Integrate and Dump ("I&D") circuits **180** and **182,** the output of which can then be sampled sequentially using switch **184** that is connected via line **185** to the analog to digital ("A/D") converter **188.** The values acquired by the A/D **188** are communicated to the DSP/uC **152** via data bus **190**. It should be noted that the output impedance of the low pass filters **164** and **166** should be sufficiently large in order that the DC components

generated by the balanced mixers in **138** be measurable. For every frequency k, two independent measurements are performed by setting the RAM contents so that the DACs **168** and **170** generate in the first measurement the values

$$B_1(t)=a\ \cos(2\pi kt/M)\ ;\ B_2(t)=a\ \sin(2\pi kt/M)$$

and in the second measurement the values

$$B_1(t)=a\ \sin(2\pi kt/M)\ ;\ B_2(t)=a\ \cos(2\pi kt/M)\ .$$

where a is an appropriate amplitude scale factor adapted to the DACs dynamic range.

The measured scalars obtained in I&D's **180** and **182** during the first measurement (acquired sequentially using both states of switch **184**) are designated here by x_k, v_k respectively, while those similarly obtained in the second measurement, are designated as y_k, u_k respectively. It can be shown that the first measurement corresponds to analysis of the upper sideband, a step hereinafter denoted by **AN+**, while the second measurement corresponds to an analysis of the lower sideband channel, denoted here by **AN-**. Hence, four parameters for each frequency k are obtained, thereby enabling the determination of the incoming signal phasors for both the upper (+) and lower (-) sideband channels. These are given by a relation related to the inverse of Equation (1), namely:

$$\begin{bmatrix} AI_k^- \\ AQ_k^- \\ AI_k^+ \\ AQ_k^+ \end{bmatrix} = \frac{1}{2}\begin{bmatrix} 0 & 1 & 1 & 0 \\ 1 & 0 & 0 & -1 \\ 0 & -1 & 1 & 0 \\ 1 & 0 & 0 & 1 \end{bmatrix}\begin{bmatrix} x_k \\ y_k \\ u_k \\ v_k \end{bmatrix} \qquad (2)$$

The phasor values obtained in Equation (2) during the measurement periods **AN+** and **AN-**, hereinafter termed analysis periods, are used by the DSP/uC **152** for computing and adjusting the content of RAMs **171** and **172** for the injection period in a manner to be further discussed below, so as to provide for the precise phase and amplitude relationship required of the injected signals during the HBI based on the

received phasor analysis. For the center carrier analysis ($k=0$), the system degenerates into a single conversion receiver, in which one applies 0 values for $B_1(t)$ and $B_2(t)$ and with a single measurement, one directly obtains the two quadrature component estimates at the I&D's **180** and **182** outputs.

In order to perform the injection and analysis in coordination with the group video timing, a vertical frame reference data sequence is inserted in the VBI of all, or at least one, of the channels within the group. The apparatus of Figure 3 provides for vertical frame synchronization by the use of data matched filter **194** and sync and data detector **196**, resetting the timing chain circuitry **210** via sync detect line **202**. The timing chain circuitry **210** provides all timing signals (t1 to t7) controlling the RF switches, the I&D's and the A/D strobe. The address generator **150** is provided with the proper reset and preload signals from the timing chain circuit **210** so as to scan the proper RAM memory locations containing the appropriate data records for the synthesis of the injection signals and the analysis signals as required. As explained above, based on the reciprocity property of the broadband generator **136**, it also serves as a broadband quadrature synchronous receiver which is used for phasor analysis and data and sync acquisition on selected channels. A particular channel can therefore be received by the additive combination of the quadrature signals at **161** and **162** via the summing resistors **193** and **192** that are connected to the matched filter **194**. The channel that is being received can be determined by the proper choice of the sinusoidal components being generated by the DAC pair **168,170** so that the appropriate sinusoids are applied in quadrature at **161** and **162** respectively, essentially in the same manner as that described for the analysis measurements AN+ or AN-, depending whether an upper side or lower side channel is being received. For example, if data in the VBI of an upper sided channel corresponding to frequency index k is being received, the RAM values of $B_1(t)$ and $B_2(t)$ are set to

$$B_1(t)=a\ \cos(\frac{2\pi}{M}\quad t)\ ;\ B_2(t)=a\ \sin(\frac{2\pi kt}{M}+\theta)\quad ;$$

where θ is the phase angle required to receive this channel in-phase based on the measured components AI_k^+ and AQ_k^+. It can be shown that the above trigonometric expressions for the RAM values required in the receive mode, correspond to the values which would be obtained according to Equation (1) with only one term in the sum (corresponding to frequency index k), wherein I_k^- and Q_k^- are set to zero. Stated another way, the RAM contents required in order to receive a selected channel is the same as that which is required to generate a single carrier, in-phase coherence with that channel's received carrier.

With the proper RAM signals set as described above, data inserted in the VBI of any channel within the group can be selectively detected by the matched filter **194** and the data detector **196**. This detected data stream is used to address the subscriber units in a manner known in the art of addressable subscription control. The identity of the specific channels authorized for descrambling and those denied from each subscriber will then be communicated to the microcontroller section of **152** of each subscriber. Based on this subscription configuration information, the DSP/uC portion of **152** calculates the specific linear combination required for $B_1(t)$ and $B_2(t)$ in accordance with Equation (1) for each frequency k.

To see this in the specific example of Figure 2, one first designates the lower sided channel corresponding to frequency index k as "-K" and the upper sided channel of that frequency index as "+K". Accordingly, note that Figure 2 contains nine channels with the designations -4,-3,-2,-1,0,+1,+2,+3,+4. Given this channel designation method, Figure 2b indicates a specific subscription configuration which calls for DESCRAMBLING (in phase injection) of Channels -3,-1 and +3, shown as injected signals **50**, and further DENIAL (out-of-phase injection) of Channels -2 and +2, shown as injected signals **52** and finally NO PROCESSING (no injection) on Channels -4,0,+1 and +4, shown at **54**. It is assumed that the values of AI_k^+,AQ_k^+,AI_k^- and AQ_k^- for the channels in need of injection have been obtained by the Analysis measurement sessions **AN+** and **AN-** as described above and in accordance with Equation (2). Based on this subscription configuration, the following vector entries should be used for each k in the right hand side of Equation (1) in order to generate the required subscription configuration of Figure 2:

k=0: $\begin{bmatrix}0\\0\\0\\0\end{bmatrix}$; k=1: $\begin{bmatrix}d\cdot AI_1^-\\d\cdot AQ_1^-\\0\\0\end{bmatrix}$; k=2: $\begin{bmatrix}-g\cdot AI_2^-\\-g\cdot AQ_2^-\\-g\cdot AI_2^+\\-g\cdot AQ_2^+\end{bmatrix}$; k=3: $\begin{bmatrix}d\cdot AI_3^-\\d\cdot AQ_3^-\\d\cdot AI_3^+\\d\cdot AQ_3^+\end{bmatrix}$; k=4: $\begin{bmatrix}0\\0\\0\\0\end{bmatrix}$

where d and g are appropriate positive scale factors relating the relative magnitude of the measured carrier during the analysis period and the injection phasor magnitudes required for descrambling and out-of-phase injection denial respectively.

As recalled from a previous discussion, the injected signals must be VSB modulated. This can be accomplished by providing digital samples to DACs **168** and **170** in accordance with digitally precalculated sequences which specifically yield injection pulses onset and terminations with VSB spectra. For a single carrier systems, there exists known methods of providing two quadrature components with certain relationships that yield a VSB modulated signal. A discussion of such methods can be found in the book entitled *Principles of Data Communications* by R. Lucky, J. Salz and E.J. Weldon published by McGraw-Hill, New York 1968. It can be mathematically shown that two such signals used for quadrature modulation of a carrier yielding a desired VSB pulse shaping are given by *C(t)* and *S(t)* shown in Figure 4 wherein *C(t)* is the modulating function of the in-phase component and *S(t)* is that of the quadrature component. Recall that the case of CW injection **30** of Figure 1b is characterized by a time invariant column phasor vector on the right hand side of Equation (1). Thus, for the inclusion of VSB transitions **32,33,** the time invariant phasor column in Equation (1) must be replaced by the time varying vector $R_k(t)$ given by:

$$R_k(t) = \begin{bmatrix} C(t) & S(t) & 0 & 0 \\ -S(t) & C(t) & 0 & 0 \\ 0 & 0 & C(t) & S(t) \\ 0 & 0 & -S(t) & C(t) \end{bmatrix} \begin{bmatrix} I_k^- \\ Q_k^- \\ I_k^+ \\ Q_k^+ \end{bmatrix} \qquad (3)$$

Since this expression is used in Equation (1) for each frequency k, and since $C(t)$ and $S(t)$ are not periodic functions, generally, the values of $B_1(t)$ and $B_2(t)$ will not be periodic in 12 samples, but rather a longer data record will be required for the synthesis of these spectrally shaped signals. Hence, with an appropriately increased value for the sample number M, and the numerical values for the functions $C(t)$ and $S(t)$ stored in a ROM inside the DSP/uC **152**, the RAM record values $B_1(t)$ and $B_2(t)$ are calculated by the DSP/uC **152** in accordance with Equation (1-A):

(1-A)

$$\begin{bmatrix} B_1(t) \\ B_2(t) \end{bmatrix} =$$

$$= \sum_{k=0}^{k=N} \begin{bmatrix} \cos(2\pi k\frac{t}{M}) & \sin(2\pi k\frac{t}{M}) & 0 & 0 \\ 0 & 0 & \cos(2\pi k\frac{t}{M}) & \sin(2\pi k\frac{t}{M}) \end{bmatrix} \begin{bmatrix} 0 & 1 & 0 & 1 \\ 1 & 0 & -1 & 0 \\ 1 & 0 & 1 & 0 \\ 0 & -1 & 0 & 1 \end{bmatrix} \begin{bmatrix} \cdot \\ \cdot \\ R_k(t) \\ \cdot \\ \cdot \end{bmatrix}$$

It should be further noted that a multiplicity of incrementally spaced VSB signals similar to those in Figure 2b would thus be generated on line **134**. If, for example, $N=4$, Equation (1-A) will yield 9 independent VSB modulated phasors generated with nominal phasor values determined by the selected phasor injection values for I_k^+,Q_k^+,I_k^- and Q_k^- in Equation (3), with $k = 0,1,2,3,4$. These phasor values will be set in accordance with the desired processing action on each channel, based on the subscriber's subscription status.

There exists a tradeoff between the RAM record length required for representing transients **32** and **33** of Figure 1b and the equivalent channel passband frequency response and transition bandwidths of the VSB spectra. In the example of the first preferred embodiment of the present invention, the transition intervals **402** and **403** in Figure 4, each correspond to a record length of 109 samples at 72 MSps for each of records **32** and **33** respectively. Figure 5 depicts the frequency response **500** of an equivalent digital filter that was computationaly applied to a square modulated pulse from

which the baseband quadrature functions *C(t)* and *S(t)* shown in Figure 4 are derived. As can be seen in Figure 5, one can achieve an acceptable VSB spectral shaping about the picture carrier with this relatively short sample sequence. Another feature provided by the digital signal processing capability of the present invention is the ability to design the functions *C(t)* and *S(t)* based on a precise computational digital filter of Figure 5, wherein specific spectral nulls are provided around the audio subcarrier frequencies **501**, thereby ensuring that the broadband injected VSB signals do not interfere with the audio portion of the signals. These strong attenuation features will in general not be required on frequencies which only affect the video portion of adjacent channels, since the resultant crosstalk effects are video frame synchronous for a contiguous group of channels and thus introduce only small transient artifacts at the edges of the HBI which are unviewable. Referring again to Figure 5, a specific provision for a spectral null at the chroma subcarrier frequency **502** of the lower adjacent channel is made so as to minimize chroma interference to the lower adjacent channel bordering the frame synchronous group of channels, should it's video timing be horizontally shifted in accordance with a video staggering method to be explained below.

Summarizing the discussions above, it is this specific ability of generating multichannel signals with essentially arbitrary amplitude and phase in video synchrony and RF coherence with incoming signals that allows one to provide simultaneous processing of a group of channels.

Because of practical limitations such as DAC clock rates and DAC dynamic range and linearity, the number of channels that can simultaneously and accurately be generated in the broadband generator **136** of Figure 3 is limited. It is therefore desirable to be able to use the broadband generator **136** in a timesharing mode, whereupon several channel groups may be processed sequentially, thereby increasing the total number of channels processed by switching the local oscillators feeding the broadband generator **136** to the center frequency channels of each group. This operation mode is generally possible in this video application since the required processing (and injection) time per line is limited to the HBI which is less than one forth of the total horizontal line time. Thus, at the headend, groups of channels are video synchronized in a staggered manner so that their HBI do not overlap. This is shown schematically in Figure 6a. Time

intervals **600** provide a guard interval, during which no injection is required, for switching the center carrier local oscillators among the four channel groups A through D (switch **144** in Figure 3) and the appropriate group band pass filter in filter bank **174** using switch **122**. Two possible frequency arrangements for channel assignments for each group are shown in Figures 6b and 6c. The first preferred embodiment of Figure 3 utilizes the frequency block grouping of Figure 6b, since it requires lower DAC speeds for the same number of channels per group. However, at the expense of faster DACs and RAMs, the frequency interlaced grouping of Figure 6c offers potential savings in the filter bank, since in this case only a single (and wider) harmonic filter is required. In the preferred embodiment of channel frequency grouping of Figure 6b, the local oscillator of frequency Fa **602** is selected by switch **144** during the interval labeled as GROUP A in Figure 6a. Subsequently, switch **144** is similarly moved to select the local oscillator corresponding tc the center channel of channel GROUP B, at frequency Fb **604** and so on for groups C and D, to fill out the rest of the horizontal video time, sequentially dwelling on frequencies Fc **606** and Fd **608** for equal periods of time corresponding to the HBI duration. In this manner, the same broadband generator performs the required processing for each channel in the four groups by coherent injection during the HBI of every channel.

Figure 7 shows the method of digital generation of the injected waveforms during one horizontal video line for all channel groups. As can be seen, in each channel group period, there are three RAM segments each corresponding to sequence of samples required for the respective time segments **32,30,33** of Figure 1b. The address value schedule generated by the address generator **150** is shown by **700**. The RAM address value at each time is shown on the vertical axis range **702**. It is generated by the address generator **150** and supplied to B1 and B2 RAMs **171** and **172** simultaneously via RAM address bus **156**. The address generator **150** is advanced by the clock signal on line **148**, said clock signal frequency is preferably an integral multiple of the channel incremental frequency which is 6 MHz in North America. Hence, in the example of the present invention the clock frequency is 72 MHz.

The numerical values stored in B1 and B2 RAMs, that are fed to DACs **168** and **170**, are prepared by the DSP/uC **152** based on calculations specified in Equation (1-A)

for each channel group, thus filling in the values in the RAM VSB "rise" section **732**, CW section **730** and VSB "fall" section **733** for channel GROUP A and similarly for the other three groups.

Turning to the first embodiment of the present invention, the system of Figure 3 maintains its injection and analysis waveforms in a fixed phase and amplitude relationship to the incoming signals by closing a phasor control loop for each processed channel by making successive analysis and corrections, thereby tracking any slow relative phase or amplitude drifts in the CATV distribution system or any of the components in the subscriber unit such as power splitter **102**, directional coupler **108** or the broadband amplifier **104** or any of the components within the broadband generator **136** which may affect the relative injected phasors as compared to the incoming channel phasors.

In order to better understand the phasor tracking and calibration method of the present invention, reference is made to Figure 8. It is assumed that the phasor diagrams in Figure 8 correspond to phasors observed on line **109** using the analysis method discussed above for an arbitrary channel within the controllable channel group. S_1, phasor **300**, represents the picture carrier phasor of an incoming carrier with black level transmission of 0 IRE during a line portion of the VBI for which no injection is required. At this point, an analysis period can be commenced to yield an estimate of the phasor S_1. If one assumes that the sync signal is suppressed by a baseband equivalent shift of 70 IRE, one may wish to measure a portion of the signal that provides identical suppression but at a line portion which does not require reconstruction by injection, hence making available the broadband generator as a phasor analyzer. This may be done by modifying the video signal at the headend to have a 70 IRE gray pedestal in a line portion of the VBI which otherwise would be at 0 IRE blanking level. This situation is depicted by S_2, (phasor **302**) suppressed by the 70 IRE pedestal corresponding to an RF amplitude amount **303**. In theory, an analysis of S_1 and S_2 should suffice to determine their phasor difference so that it may be generated and injected by the broadband generator **136**. The difficulty with this approach stems from the fact that the signal paths for the measured signals and the injected signals are unequal and thus no direct measurement of the injected phasor as it is imparted on line **109** is available. This

primarily stems from the fact that during the injection period, the generator must generate a plurality of signals which cannot be useful for phasor analysis and vice versa. That is, only one function can be achieved at a time. It would be preferable to measure S, the combination of the incoming phasor S_2 and the injected phasor S_{inj} which would then be compared to S_1. Using these two measurements, one then derives the error phasor ΔS based on the vector equation

$$\Delta S = S_1 - S \qquad (4)$$

From the above equation, a phasor increment is calculated by DSP/uC **152** and accumulated so as to modify the RAM records for injection. This process is constructed iteratively so convergence to phasor matching is effected.

The assumption made above regarding the measurability of S is that both an injection and analysis sources exist simultaneously. It is possible to obtain a measure for S in a sequential manner by taking advantage of sufficient SAW filter delay in filter bank **174**. The method is based on an injection of the required composite RF signal during a short pulse and the subsequent reception of the delayed version of the pulse while the broadband generator becomes an analyzer. This is done by manipulating RF switches **116** and **126** during a full line of the VBI wherein a 70 IRE pedestal is inserted at the headend. In the first preferred embodiment, such calibration and analysis is performed during video lines 17 and 280. Figure 9 shows the pulsed injection and analysis timing as well as the state schedule of RF switches **116** and **126**. Figure 10 shows the related RAM addressing schedule in which injection sectors designated by Rec. A,B,C and analysis sectors **AN+** and **AN-** are periodically alternating through the control of the address generator **150**. The samples obtained by the A/D **188** in this process may be averaged in the DSP/uC **152** in order to provide additional precision for the phasor estimates thus derived. The procedure described above can be employed sequentially on all channels for which injection is required.

The above paragraph described the measurement of the combined phasor S of Figure 8. In order to perform the adjustment in accordance with Equation (4), S_1 may be obtained freely during much of the 0 IRE blanking level of the VBI. In this case, multiple channels and groups can be processed within one VBI frame. This is simply

done by subjecting the system to sequential CW analysis sessions AN+ and AN- as shown for multiple groups in Figure 11.

The overall video timing schedule of the first embodiment of the present invention is shown in Figure 12. As can be seen, the periods for injection, analysis, sync and data detection are non-overlapping and thus provide an efficient use of the broadband generator. Upon completion of each video field or frame, correction to all injected phasors can be made in accordance with increments **ΔS** yielding for all frequencies the following increments for $B_1(t)$ and $B_2(t)$:

$$\begin{bmatrix} \Delta B_1(t) \\ \Delta B_2(t) \end{bmatrix} = \sum_{k=0}^{k=N} \begin{bmatrix} \cos(2\pi k\frac{t}{M}) & \sin(2\pi k\frac{t}{M}) & 0 & 0 \\ 0 & 0 & \cos(2\pi k\frac{t}{M}) & \sin(2\pi k\frac{t}{M}) \end{bmatrix} \begin{bmatrix} 0 & 1 & 0 & 1 \\ 1 & 0 & -1 & 0 \\ 1 & 0 & 1 & 0 \\ 0 & -1 & 0 & 1 \end{bmatrix} \begin{bmatrix} \Delta I_k^- \\ \Delta Q_k^- \\ \Delta I_k^+ \\ \Delta Q_k^+ \end{bmatrix} \quad (5)$$

Figure 13 provides a schematic diagram of a CATV headend configured in accordance with the present invention in which only two groups of channels (groups B and D) contain scrambled channels to be processed by the subscriber unit of Figure 3. A comb generator supplies independent comb signals to each group, thereby allowing partial IRC operation. A more detailed view of the comb generator is provided in Figure 14. The two broadband comb generators feeding group B comb and group D comb are of similar construction to the subscriber broadband generator **136**. However, here, all carriers are constructed in a CW periodic mode. Thus, for a 72 MHz clock frequency, the address generator is essentially a modulo 12 counter providing the 6 MHz periodicity. The 72 MHz pilot signal is also fed into the CATV system in order to provide all subscriber units with synchronous clock. Note also that the master video timing supplied by the genlock reference signal is locked to the 72 MHz clock, thereby guaranteeing the rigid relationship between line rate and RF clock rate that eliminates the need to change the RF RAM records for every video line at the subscriber unit. The

relative phases of the comb signals may be changed by downloading other RAM records through data buses W_B and W_D. These phase values are practically irrelevant to the operation of the system of the present invention but it may be desirable to adjust them in order to reduce total transmission distortion as disclosed in the above cited U.S. Patent to the present inventor.

Figure 15a shows a baseband representation of an RF sync suppression HBI. Figure 15b depicts the required coherent injection signal for descrambling. Here, due to the attenuation at RF, all signal components are in need of injection including a portion of the color burst. As can be seen, six RAM records are required for generation of the picture carrier portion of the injection signal. These are **1032, 1040, 1038, 1030, 1036,** and **1033**. Although it may be possible to generate the missing burst signal by VSB generation about the picture carrier, this approach is prohibitively expensive in memory because of the poor congruence between the color subcarrier frequency and 6 MHz.

An alternative solution is based on the fact that all upper sideband color subcarriers in a frame synchronized video sources are incrementally related with 6 MHz spacing. Hence they can be generated separately with a 3.58 MHz offset local oscillator locked on the color burst of the center carrier channel as shown in the description of the second embodiment of the present invention provided by Figures 16 and 17. In a manner similar to the first embodiment, all injected phasors are measured and tracked preferably during the VBI. Figure 17 shows the arrangement for a Multiple Dwelling Unit ("MDU") subscriber group. Here, the digital broadband generator of subscriber A provides the analysis function during calibrated injection of subscriber B and vice versa.

Finally, it is worth noting that if a broadband generator is dedicated to one group during the entire video line, it can generate additional fixed injection signals of varying levels during the active video time so as to provide an additional denial security based on random "video folding". This method can be implemented by effecting gated coherent injection at the headend using some fixed set of injection values governed by cryptographic keystream control and selectively null them out (therefore descramble) at the subscriber location based on precise phasor adjustment. Because the headend injected signal can be in opposite phase, video inversion on only a sector of the screen may take

place, further frustrating any existing "pirate" descramblers. Figure 18 shows the RAM timing diagram for a descrambler utilizing the video folding descrambling and signal denial technique.

The embodiments of the invention in which an exclusive property or privilege is claimed are defined as follows:

1. A method of generating a broadband signal comprising up to $2N+1$ RF signals where N is an integer, each of the RF signals having a frequency that is related to a center frequency by a frequency difference that is a positive or negative integral multiple of a fundamental frequency, each of the RF signals having an amplitude and phase characterized by a phasor value, said method comprising the steps of:

generating a first and second baseband signal each comprising a linear combination of up to N sinusoids, each of the sinusoids having a frequency that is an integral multiple of the fundamental frequency; and

modulating in quadrature the first and second baseband signals on a local signal [146] having a frequency equal to the center frequency.

2. The method of Claim **1** wherein the RF signals are continuous wave (CW) signals.

3. The method of Claim **1** wherein the RF signals have phasor values that are each subjected to quadrature modulation.

4. The method of Claim **3** wherein the quadrature modulation of the RF signals' phasor values results in vestigial sideband modulation.

5. The method of Claim **1** further comprising the step of determining a desired linear combination of sinusoids and zero frequency terms for the first and second baseband signals to achieve a desired phasor value for each of the RF signals.

6. The method of Claim **5** wherein said step of determining further comprises the steps of:

representing the first and second baseband signals as first and second digital sample sequences $B_1(t)$ and $B_2(t)$ in accordance with

$$\begin{bmatrix} B_1(t) \\ B_2(t) \end{bmatrix} = \sum_{k=0}^{k=N} \begin{bmatrix} \cos(2\pi k F_0 t) & \sin(2\pi k F_0 t) & 0 & 0 \\ 0 & 0 & \cos(2\pi k F_0 t) & \sin(2\pi k F_0 t) \end{bmatrix} \begin{bmatrix} 0 & 1 & 0 & 1 \\ 1 & 0 & -1 & 0 \\ 1 & 0 & 1 & 0 \\ 0 & -1 & 0 & 1 \end{bmatrix} \begin{bmatrix} \cdot \\ \cdot \\ R_k(t) \\ \cdot \\ \cdot \end{bmatrix}; and$$

wherein t is sample time, F_0 is the fundamental frequency and $R_k(t)$ is the time varying vector describing the desired phasor values of the RF signals that are separated from the center frequency by a frequency difference equal to k times the fundamental frequency.

7. The method of Claim **6** wherein the digital sample sequences are provided at a sample rate that is the product of M and the fundamental frequency such that when the time t is taken as an integer sample index incrementing at the sample rate, $F_0=1/M$ where M is an integer.

8. The method of Claim **6** wherein the desired phasor values of the RF signals are modulated in accordance with independent functions.

9. The method of Claim **6** wherein:

the desired phasor values of the RF signals are defined by the equation: $R_k(t) = \begin{bmatrix} C(t) & S(t) & 0 & 0 \\ -S(t) & C(t) & 0 & 0 \\ 0 & 0 & C(t) & S(t) \\ 0 & 0 & -S(t) & C(t) \end{bmatrix} \begin{bmatrix} I_k^- \\ Q_k^- \\ I_k^+ \\ Q_k^+ \end{bmatrix}$;

$C(t)$ and $S(t)$ are functions which impose quadrature modulation of the RF signals wherein I_k^- and Q_k^- represent the in-phase and quadrature components respectively for an RF signal having a frequency lower by k fundamental frequency increments relative to the center frequency; and

I_k^+ and Q_k^+ represent the in-phase and quadrature components respectively for an RF signal having a frequency higher by k fundamental frequency increments relative to the center frequency.

10. The method of Claim **7** wherein said step of generating further comprises the steps of:

sequentially retrieving a sample pair, one from each of the first and second digital sample sequences, for each cycle of a clock having a frequency equal to the sample rate; and

converting each of the retrieved sample pairs to analog values to produce the first and second baseband signals.

11. The method of Claim **10** wherein the clock is phase-locked to a pilot signal which is an integral multiple of the fundamental frequency.

12. The method of Claim **11** wherein the local signal is phase-locked to a channel group center carrier.

13. The method of Claims **1** or **5** further comprising the step of filtering the first and second baseband signals to limit their spectral energy above the frequency of their highest frequency sinusoidal component.

14. The method of Claim **10** wherein the frequency of said clock signal is greater than the product of $2N$ and the fundamental frequency.

15. An apparatus for generating a broadband signal comprising up to $2N+1$ RF signals where N is an integer, each of the RF signals having a frequency separated from a center frequency by a frequency difference that is a positive or negative integral multiple of a fundamental frequency, said apparatus comprising:

means for generating digital samples representing first and second baseband signals, each of the baseband signals comprising a linear combination of up to N sinusoids, each of the sinusoids having a frequency that is an integral multiple of the fundamental frequency;

means, coupled to said means for generating digital samples, for converting the digital samples to the first and second baseband signals;

means for generating a clock signal coupled to said means for generating digital samples and to said means for converting;

means for generating a local signal having a frequency equal to the center frequency; and

means, coupled to the first and second baseband signals and the local signal, for modulating in quadrature the first and second baseband signals on the local signal to produce the broadband signal.

16. The apparatus of Claim **15** wherein the frequency of said clock signal is greater than the product of $2N$ and the fundamental frequency.

17. The apparatus of Claim **15** wherein the RF signals are continuous wave (CW) signals.

18. The apparatus of Claim **15** wherein the RF signals have phasor values that are each subjected to quadrature modulation.

19. The apparatus of Claim **18** wherein the quadrature modulation of the RF signals' phasor values results in vestigial sideband modulation.

20. The method of Claim **5** or **6** wherein the zero frequency terms of the desired linear combination have amplitudes of zero.

FIG. 1a

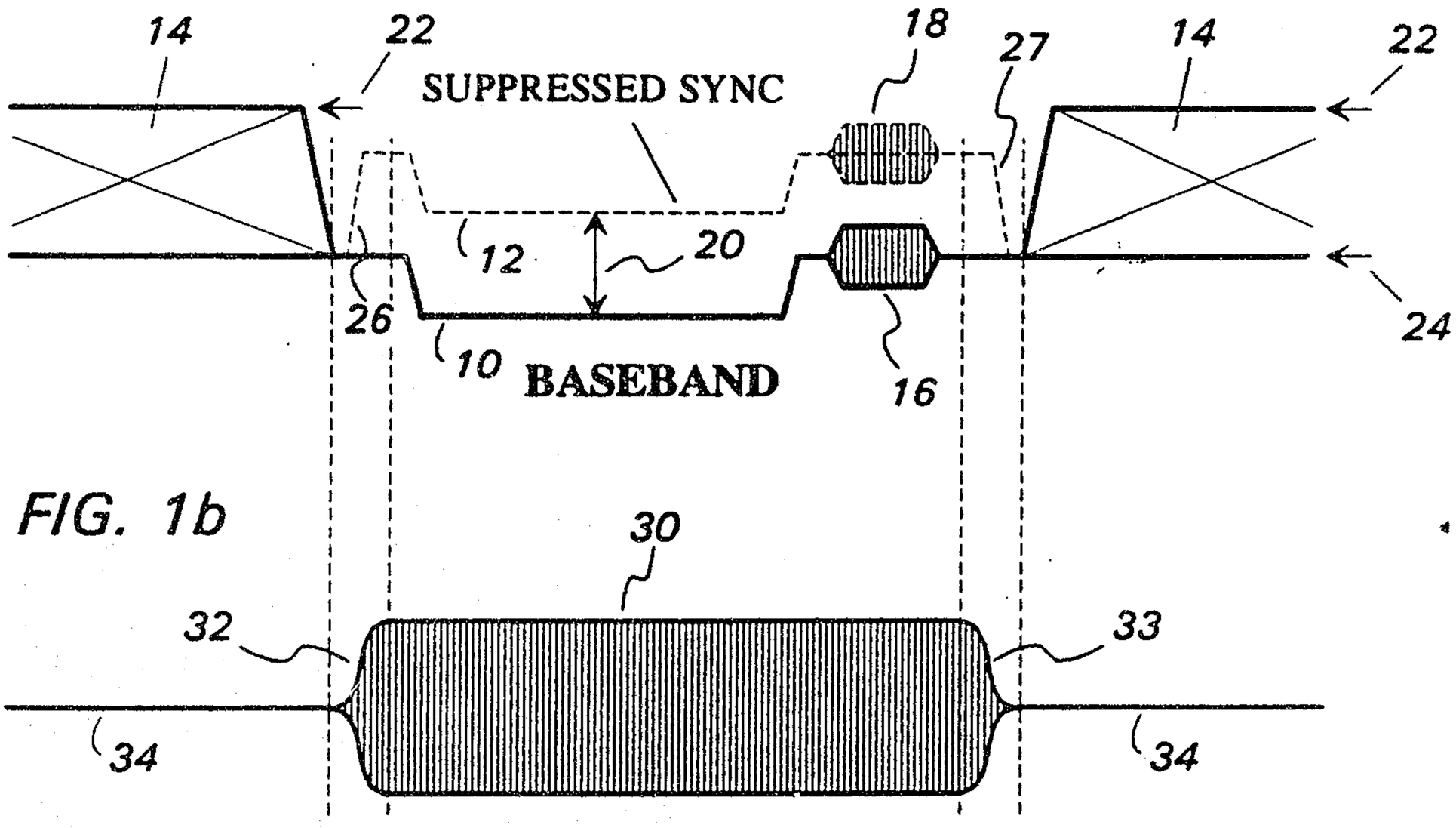


FIG. 1b

BASEBAND SYNC SUPPRESSION

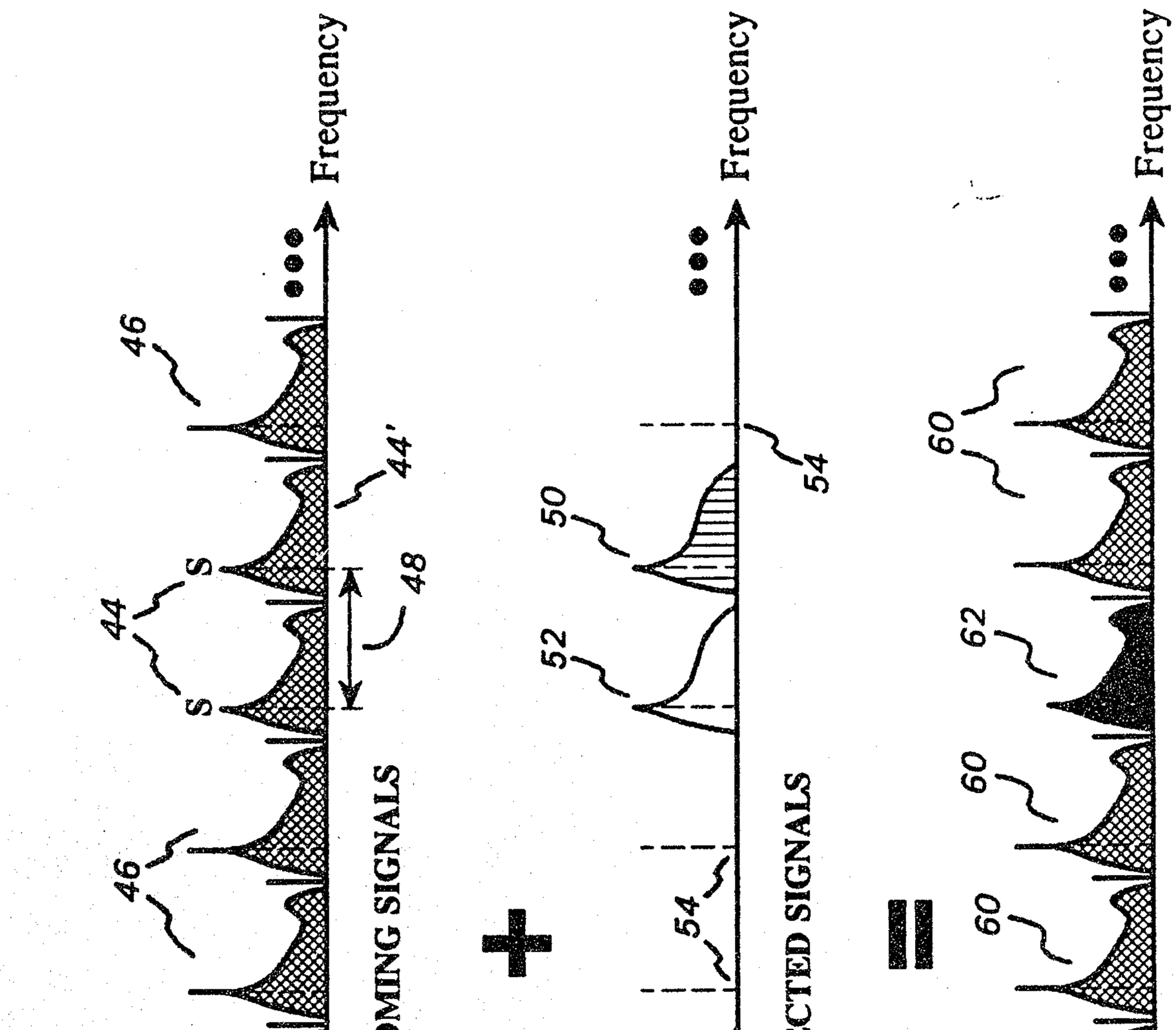
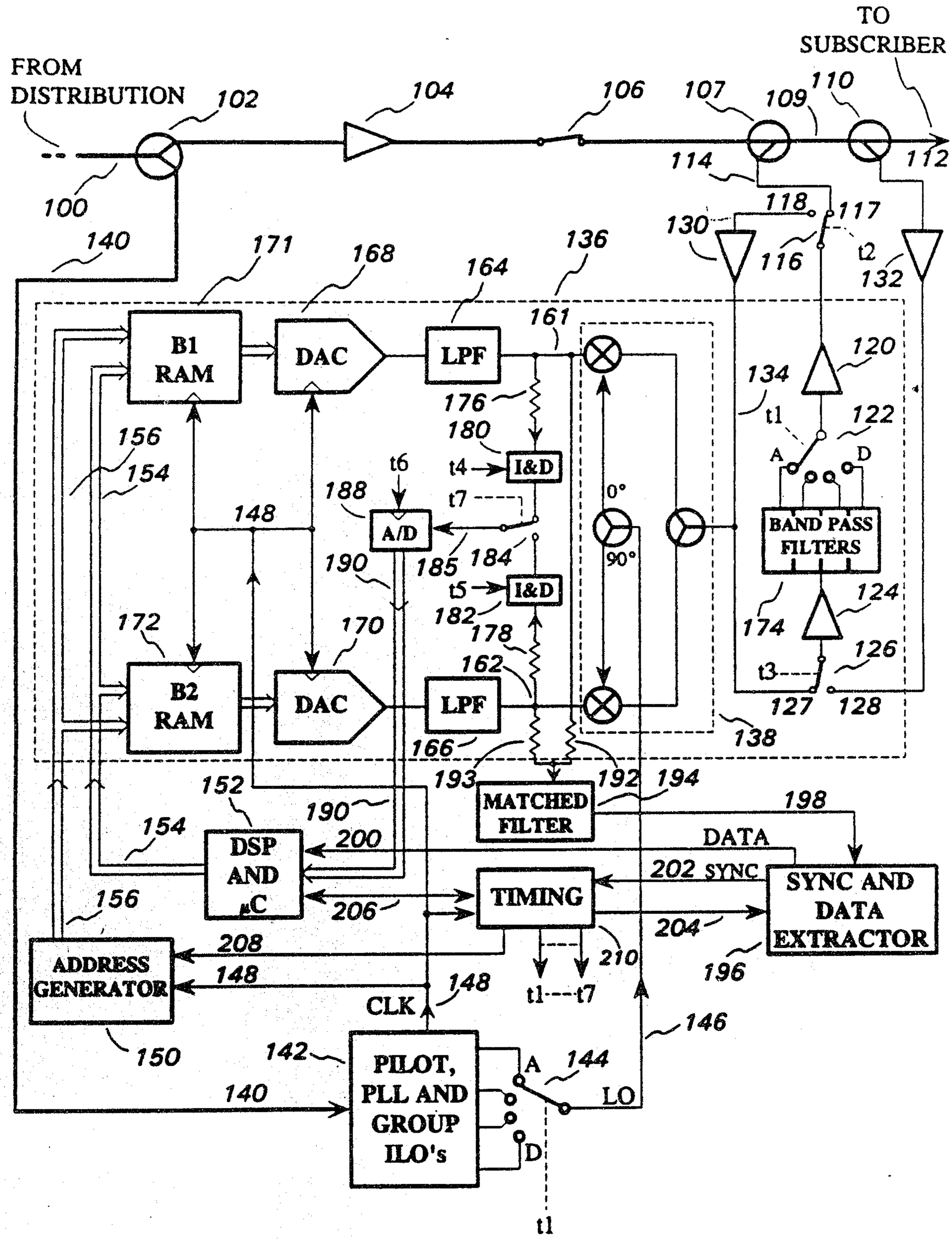


FIG. 2a FIG. 2b FIG. 2c

FIG. 3


FROM DISTRIBUTION
TO SUBSCRIBER
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102
104
106
107
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117
118
120
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128
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136
138
140
142
144
146
148
150
152
154
156
161
162
164
166
168
170
171
172
174
176
178
180
182
184
185
188
190
192
193
194
198
200
202
204
206
208
210
B1 RAM
B2 RAM
DAC
DAC
LPF
LPF
I&D
I&D
A/D
0°
90°
BAND PASS FILTERS
MATCHED FILTER
DSP AND µC
TIMING
SYNC AND DATA EXTRACTOR
ADDRESS GENERATOR
PILOT, PLL AND GROUP ILO's
DATA
SYNC
CLK
A
D
LO
t1
t2
t3
t4
t5
t6
t7

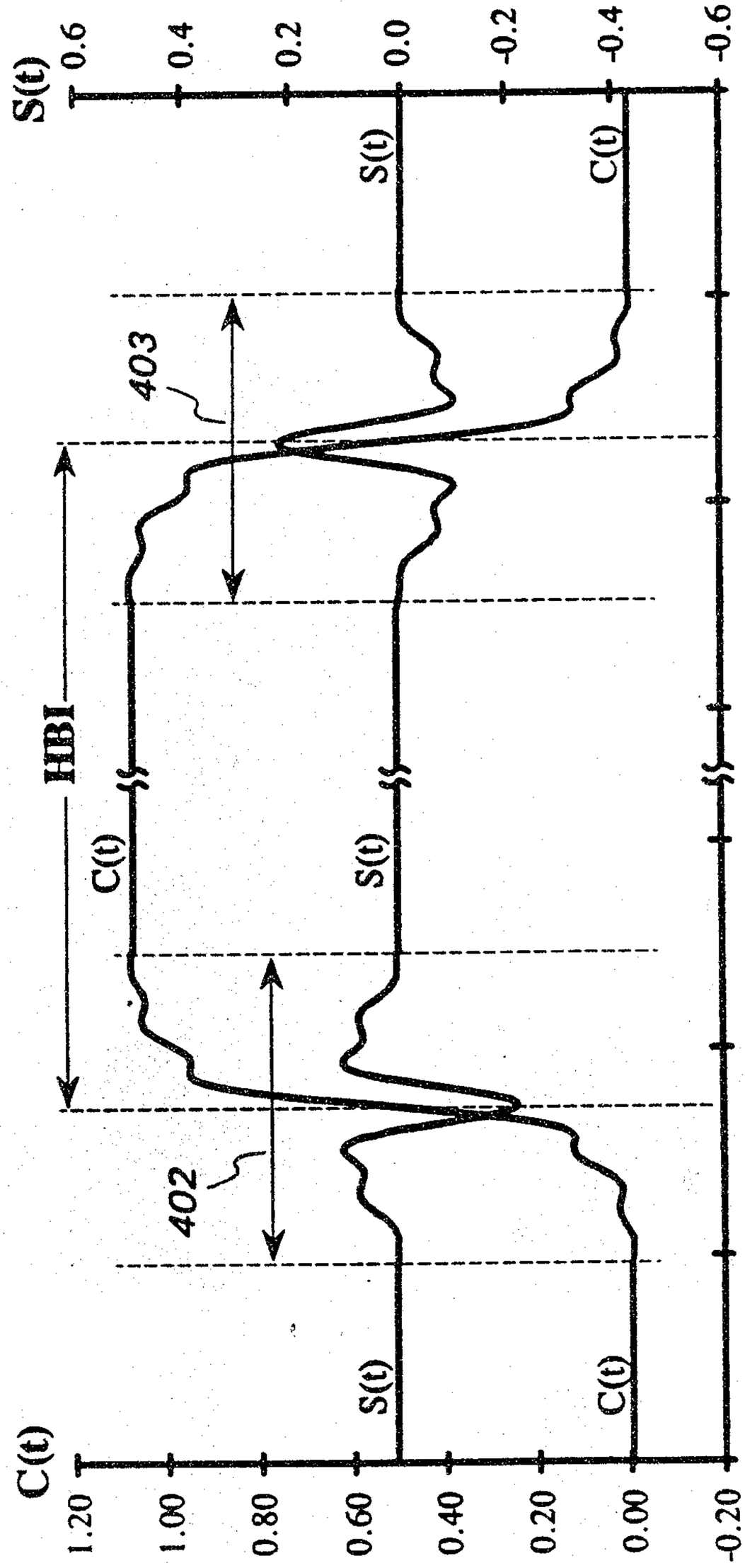

S(t)
0.6
0.4
0.2
0.0
-0.2
-0.4
-0.6
S(t)
C(t)
403
HBI
C(t)
S(t)
402
TIME
C(t)
1.20
1.00
0.80
0.60
0.40
0.20
0.00
-0.20
S(t)
C(t)

FIG. 4

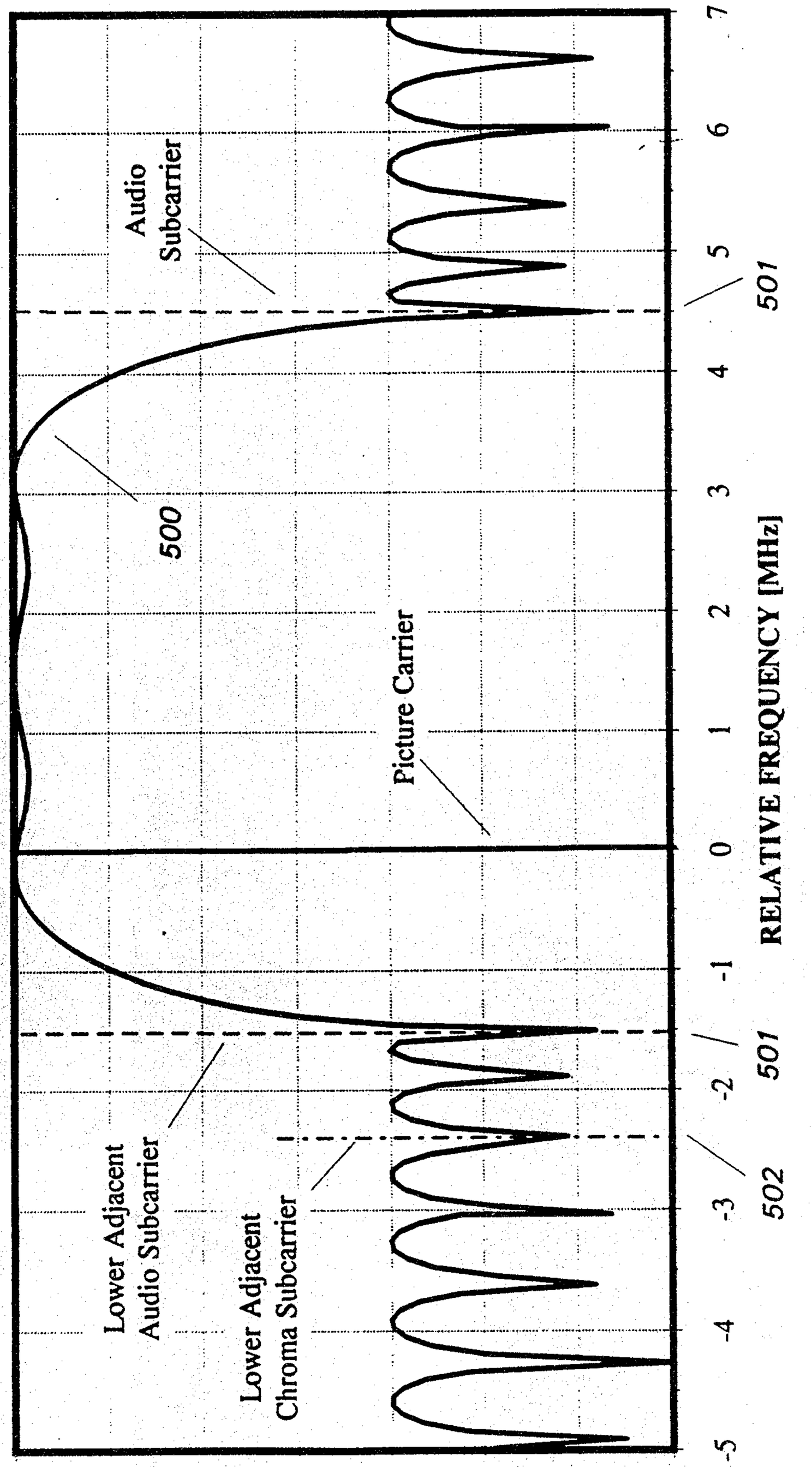
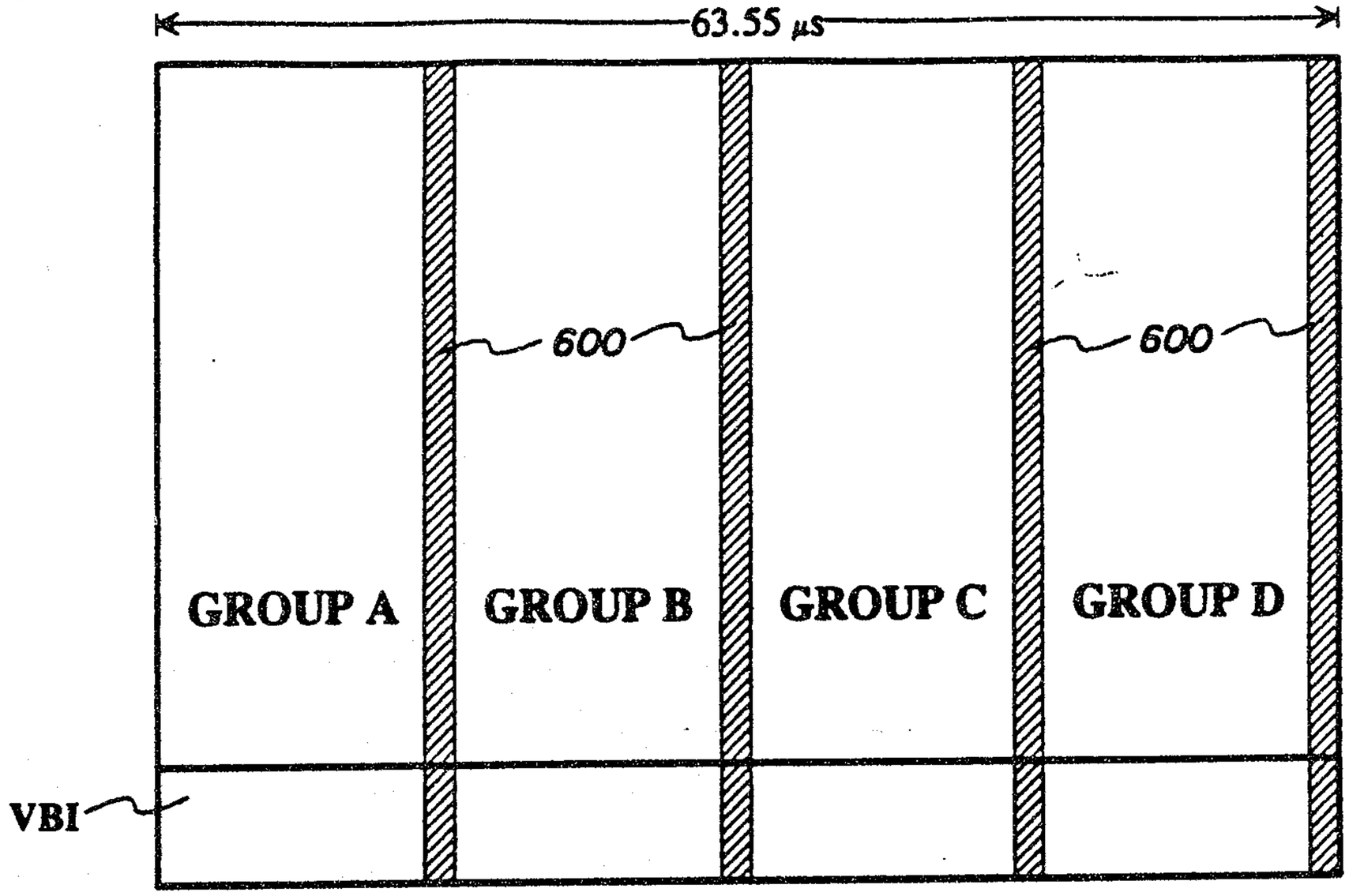

Audio Subcarrier
Picture Carrier
Lower Adjacent Audio Subcarrier
Lower Adjacent Chroma Subcarrier
500
501
502
RELATIVE FREQUENCY [MHz]
RESPONSE [dB]
0
-10
-20
-30
-40
-50
-60
-70
7
6
5
4
3
2
1
0
-1
-2
-3
-4
-5

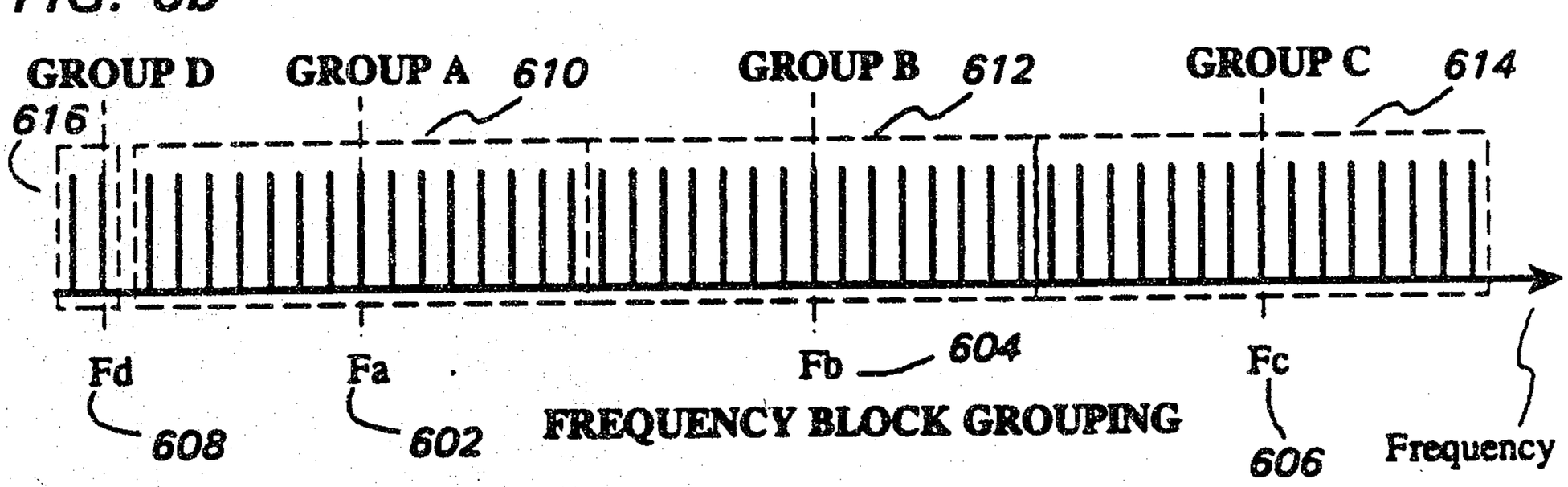
FIG. 5

FIG. 6a



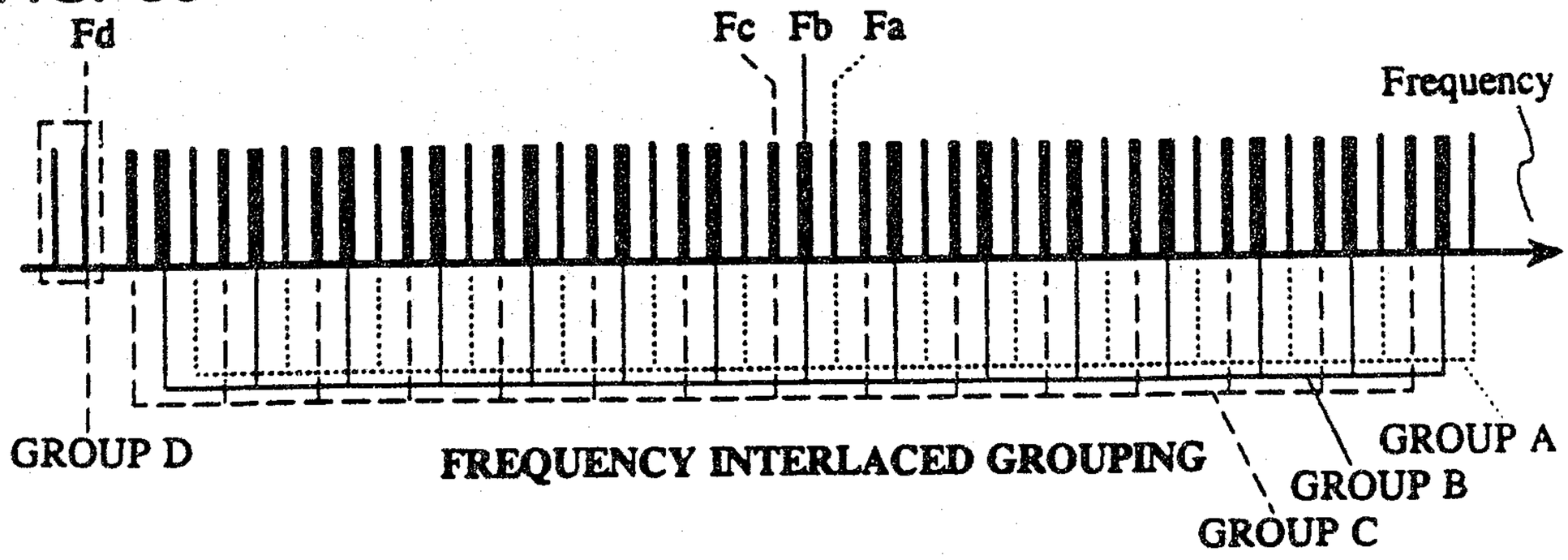
STAGGERED GROUPING IN TIME DOMAIN

FIG. 6b



FREQUENCY BLOCK GROUPING

FIG. 6c



FREQUENCY INTERLACED GROUPING

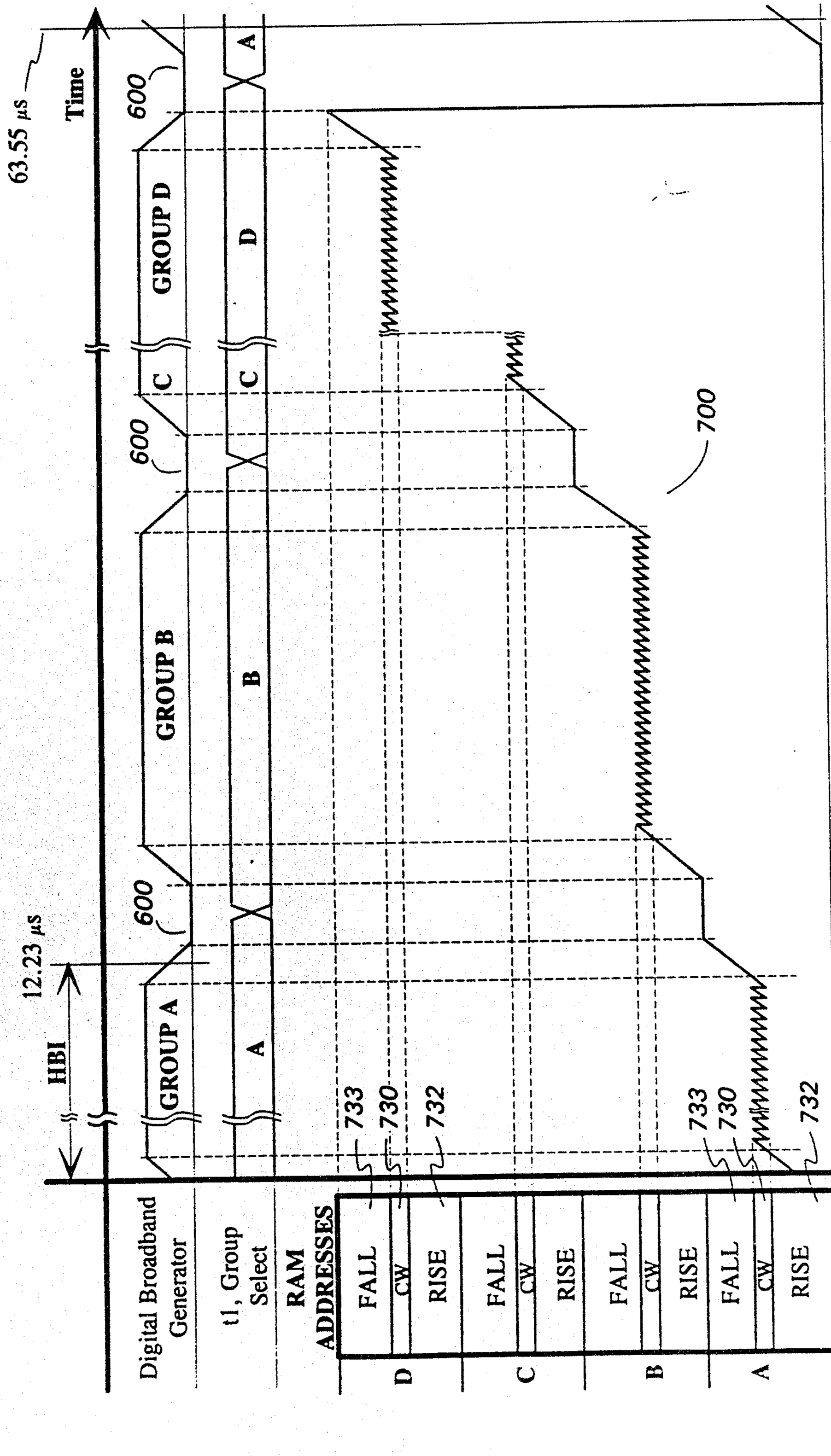
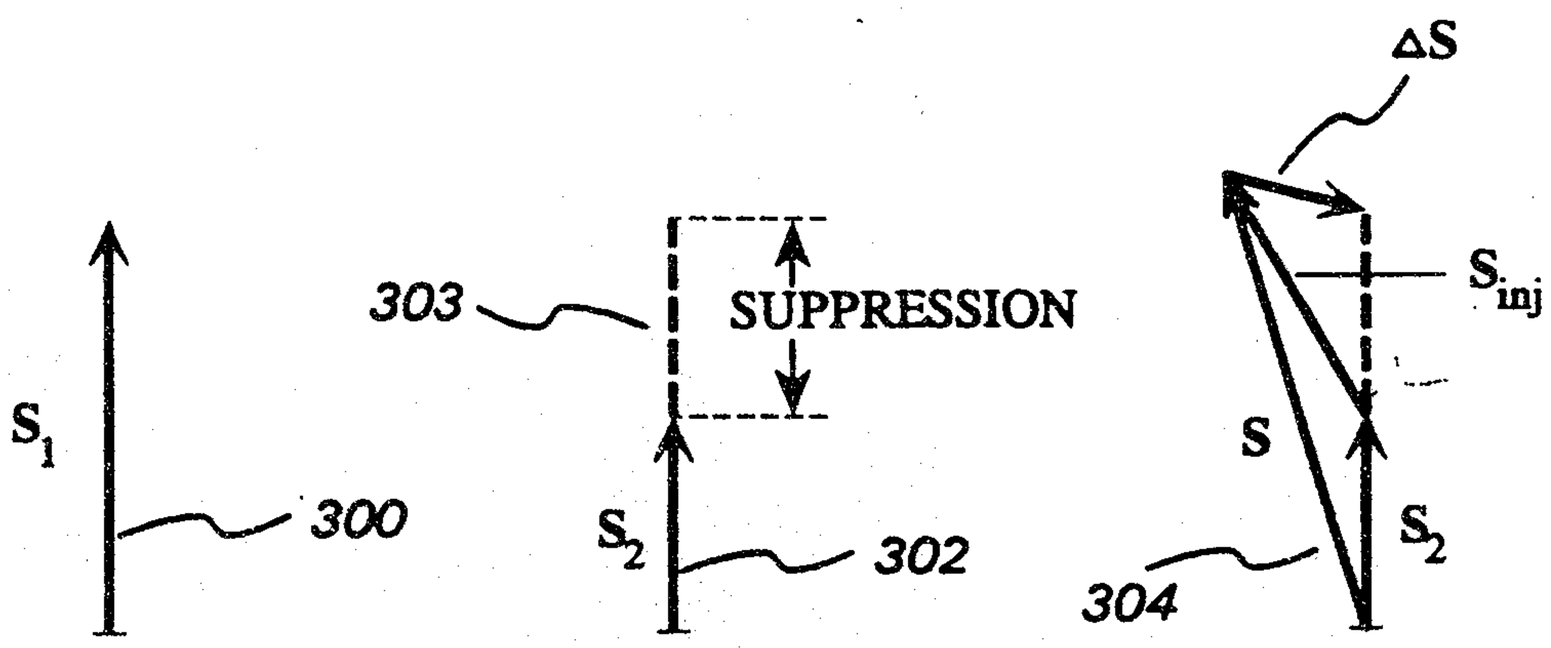

Time
63.55 μs
600
A
GROUP D
D
C
C
600
700
600
GROUP B
B
12.23 μs
HBI
600
GROUP A
A
733
730
732
733
730
732
LINES 27-260, and 290-523
Digital Broadband Generator
t1, Group Select
RAM ADDRESSES
FALL
CW
RISE
FALL
CW
RISE
FALL
CW
RISE
FALL
CW
RISE
D
C
B
A
702

FIG. 7

FIG. 8



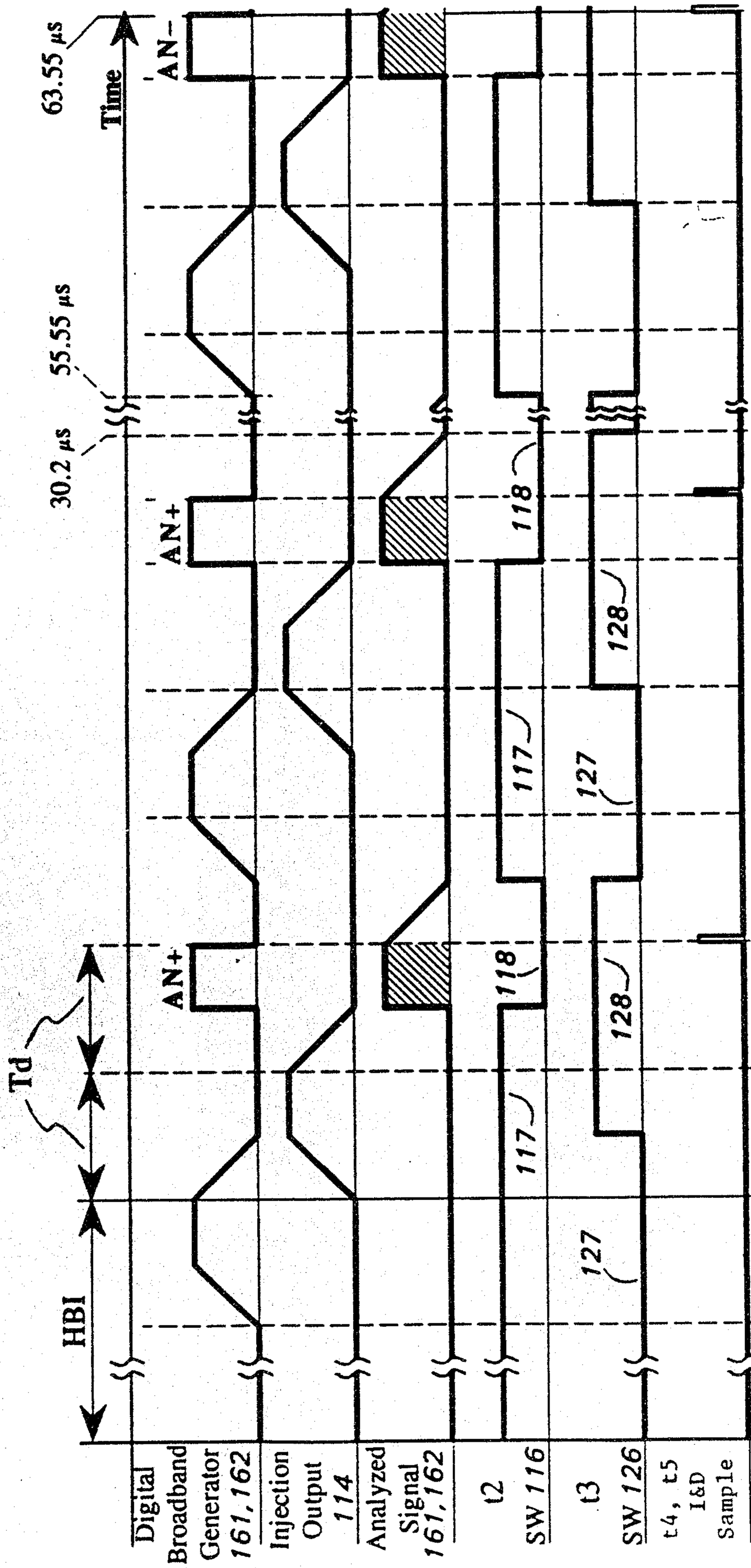


FIG. 9

LINES 17, and 280

Td = 2.67 µs

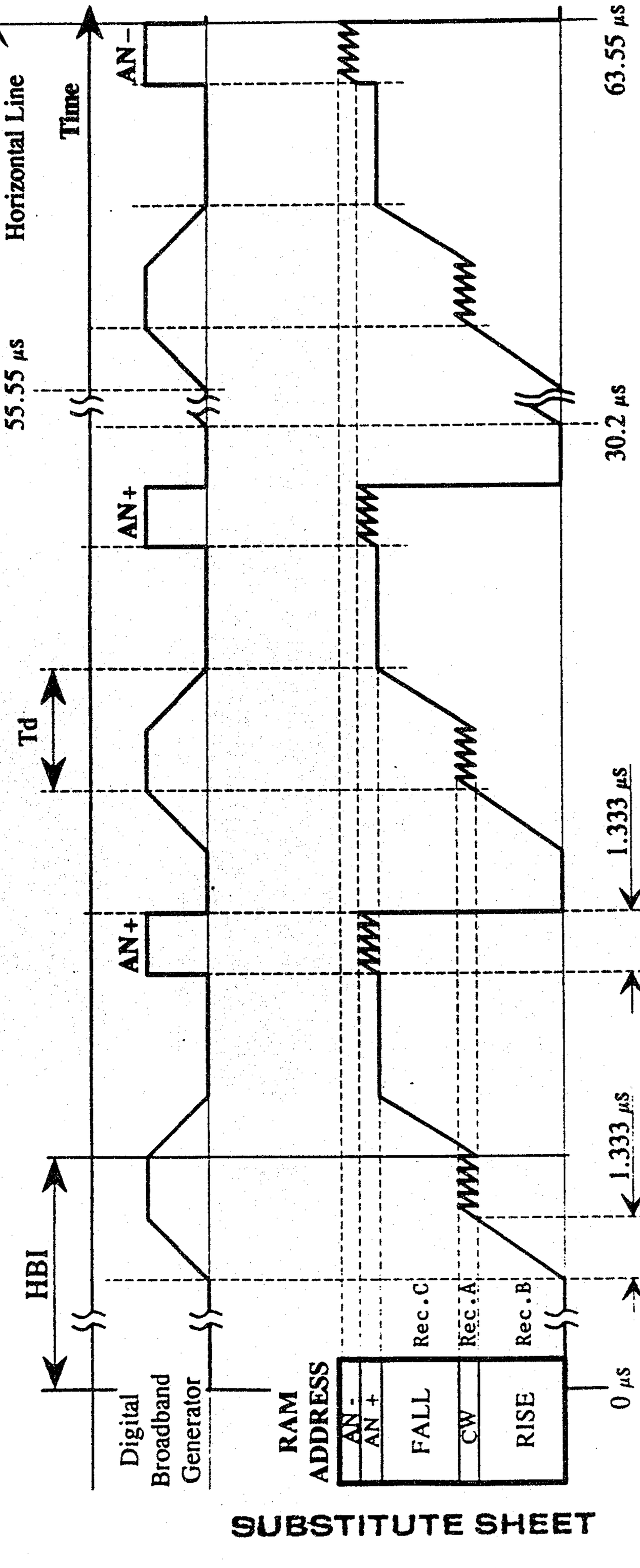


FIG. 10

LINES 17, and 280

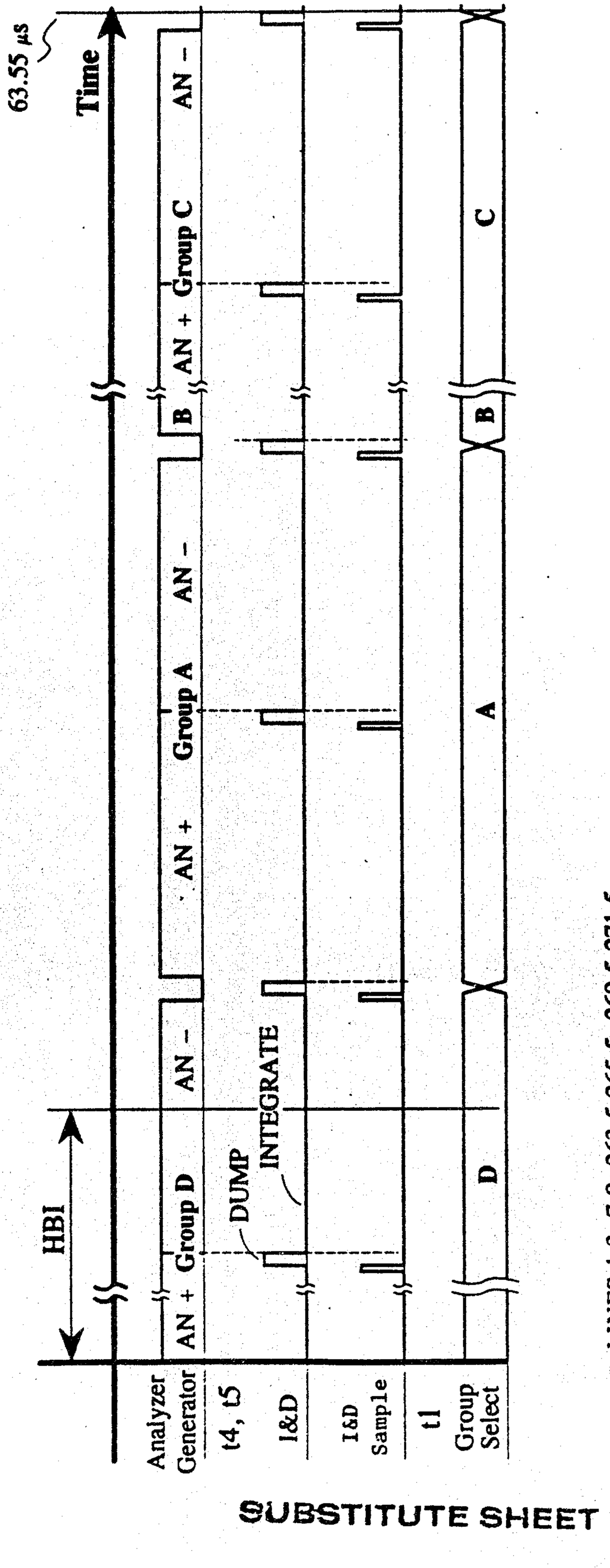

63.55 μs
Time
AN −
AN + Group C
B
AN −
AN + Group A
AN −
INTEGRATE
HBI
AN + Group D
DUMP
C
B
A
D
Analyzer Generator
t4, t5
I&D
I&D Sample
t1
Group Select
LINES 1-3, 7-9, 263.5-265.5, 269.5-271.5

FIG. 11

FIG. 12

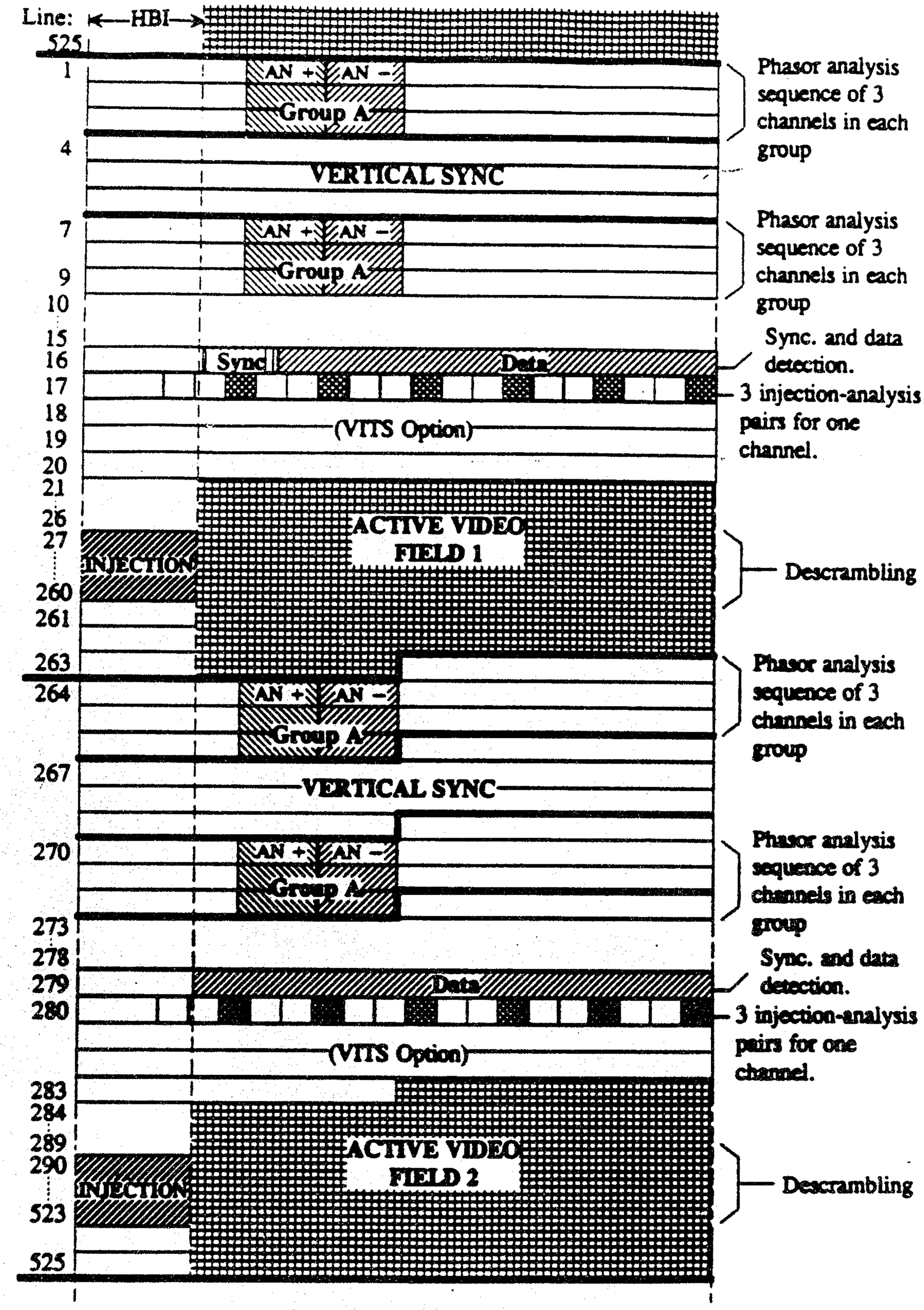

Line:
HBI
525
1
AN +
AN –
Group A
4
VERTICAL SYNC
7
9
10
15
16
Sync
Data
17
18
19
(VITS Option)
20
21
26
27
INJECTION
ACTIVE VIDEO
FIELD 1
260
261
263
264
267
270
273
278
279
280
283
284
289
290
523
525
Phasor analysis sequence of 3 channels in each group
Sync. and data detection.
3 injection-analysis pairs for one channel.
Descrambling

FIG. 13a

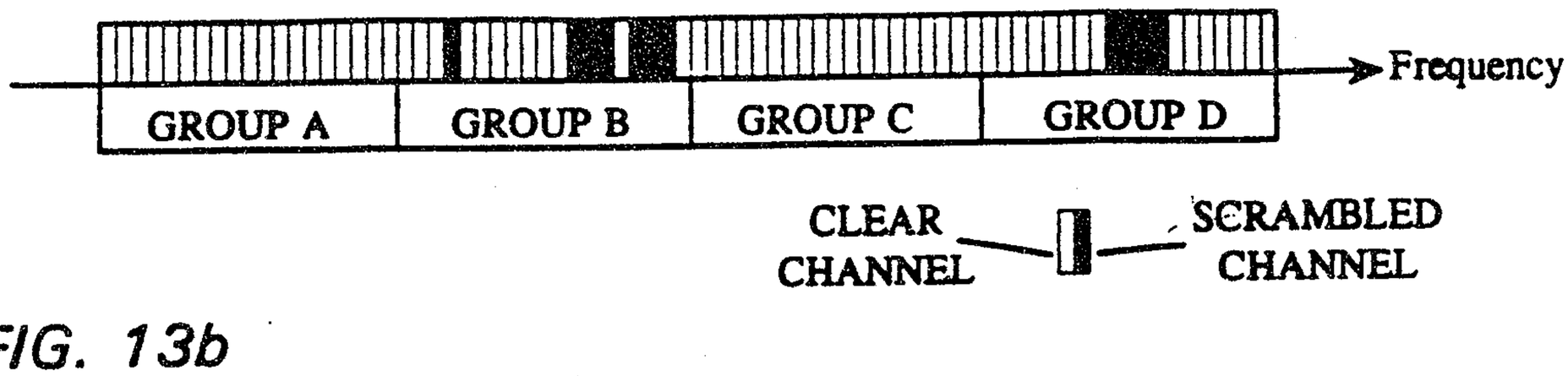


FIG. 13b

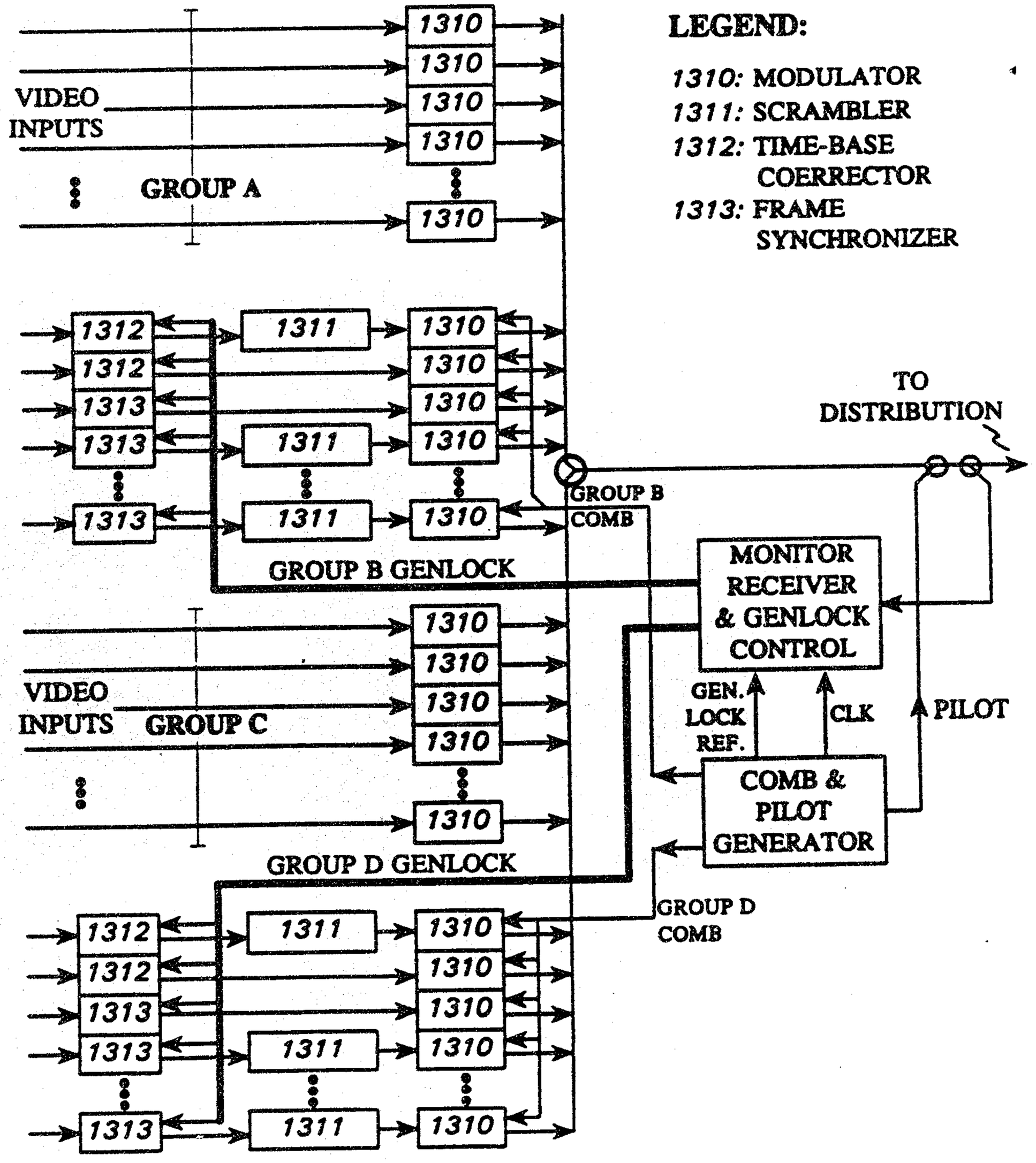


FIG. 14

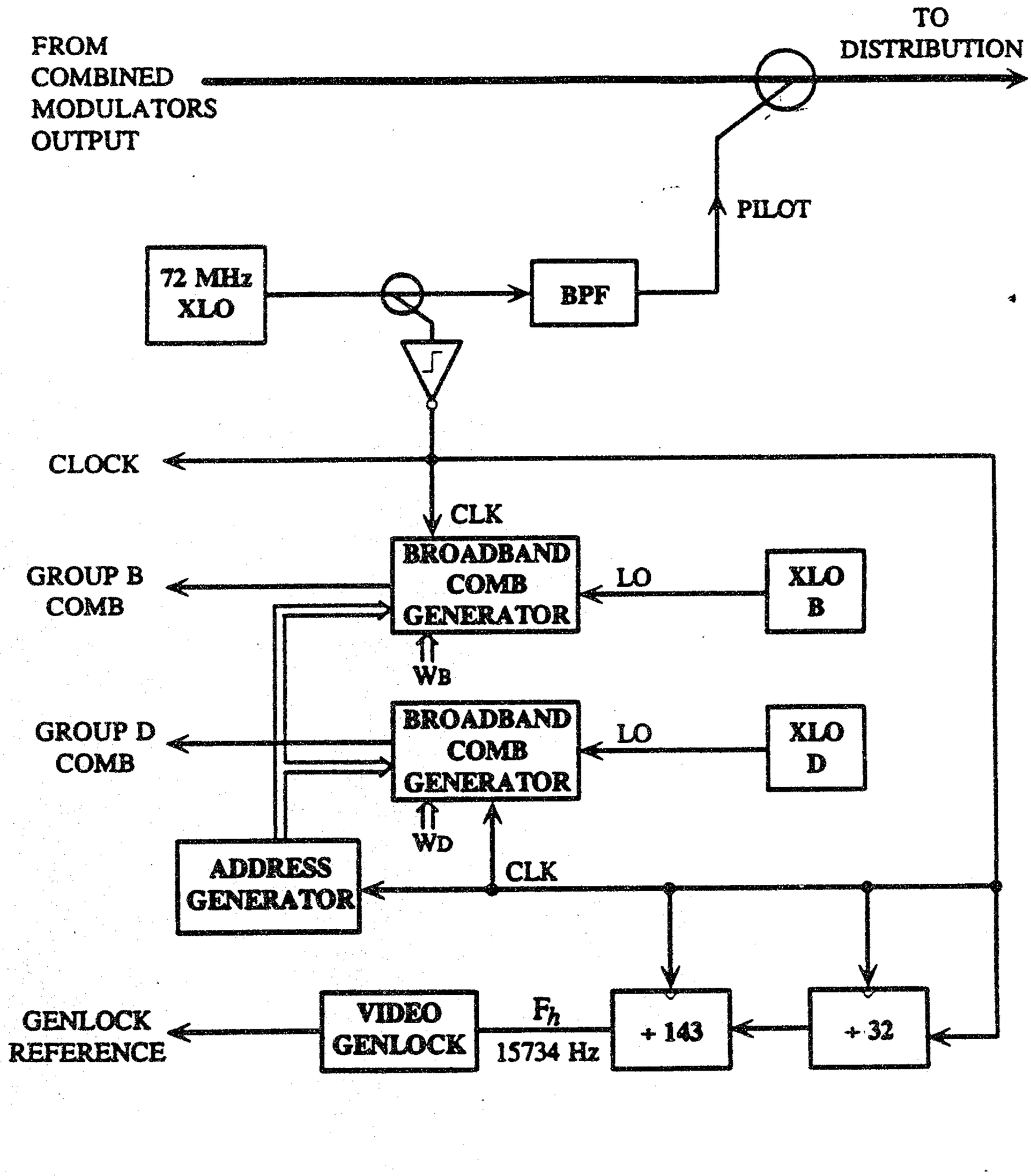

FROM COMBINED MODULATORS OUTPUT
TO DISTRIBUTION
PILOT
72 MHz XLO
BPF
CLOCK
CLK
BROADBAND COMB GENERATOR
GROUP B COMB
LO
XLO B
WB
BROADBAND COMB GENERATOR
GROUP D COMB
LO
XLO D
WD
ADDRESS GENERATOR
CLK
GENLOCK REFERENCE
VIDEO GENLOCK
Fh
15734 Hz
+ 143
+ 32

FIG. 15a

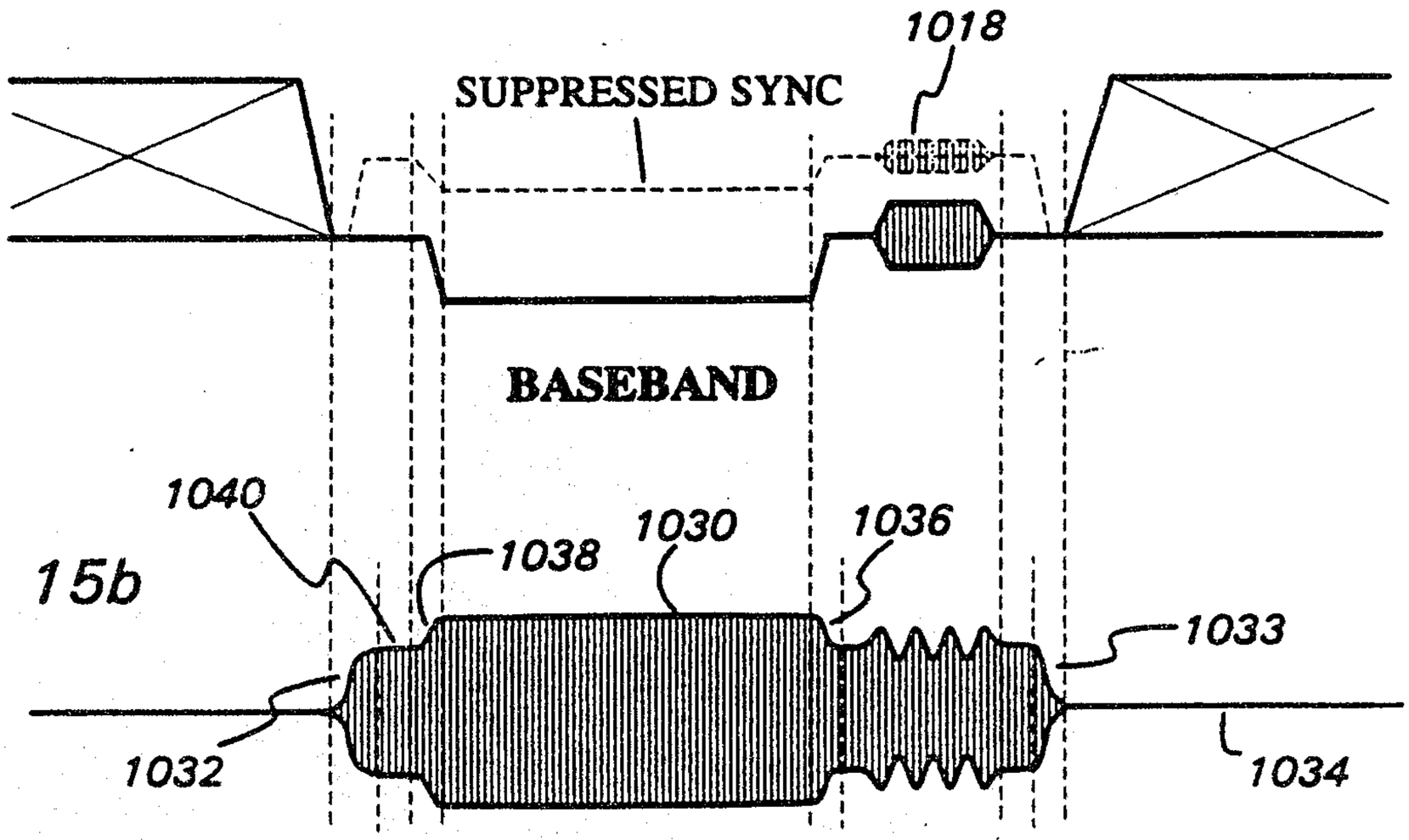


FIG. 15b

FIG. 16

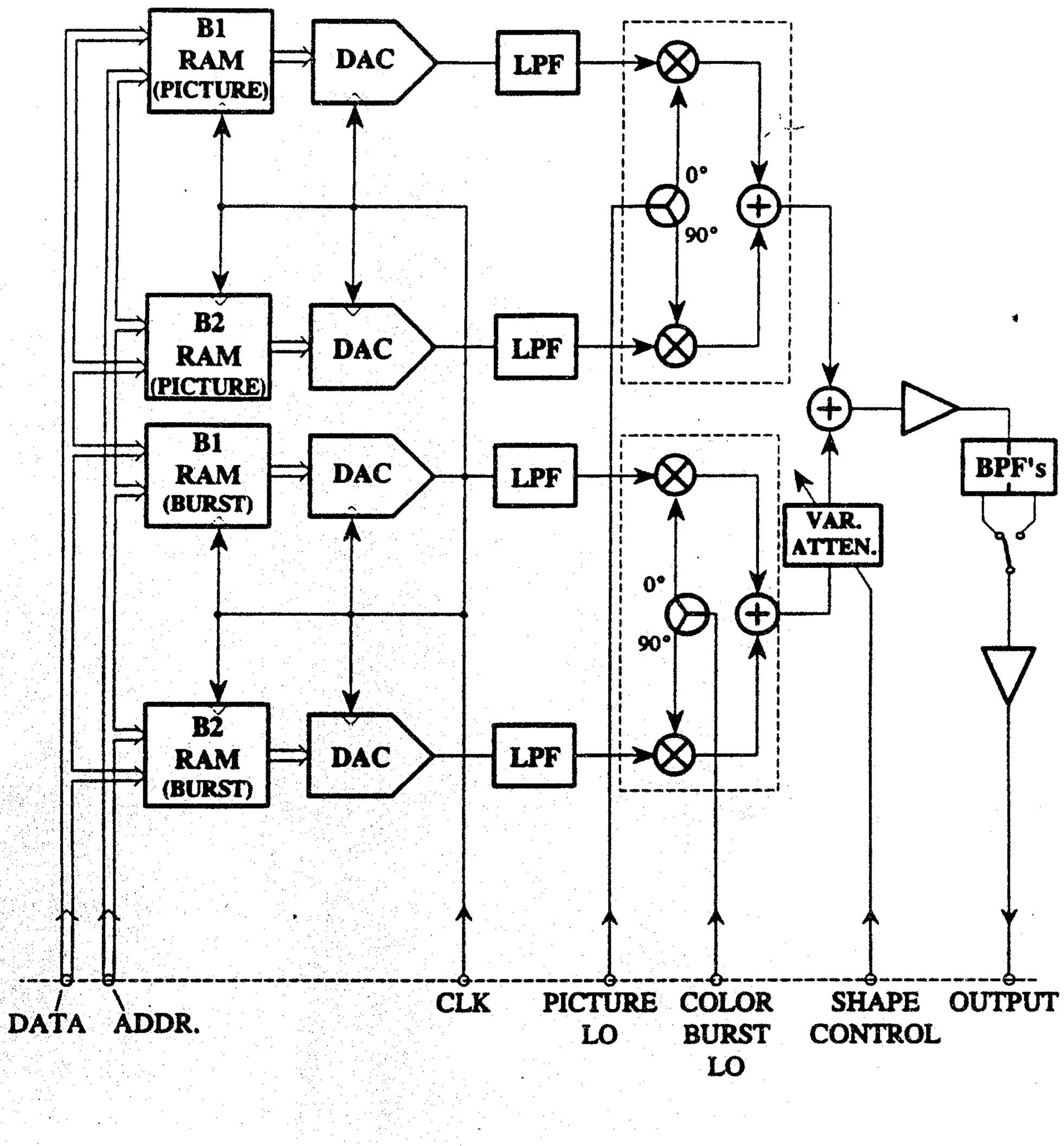
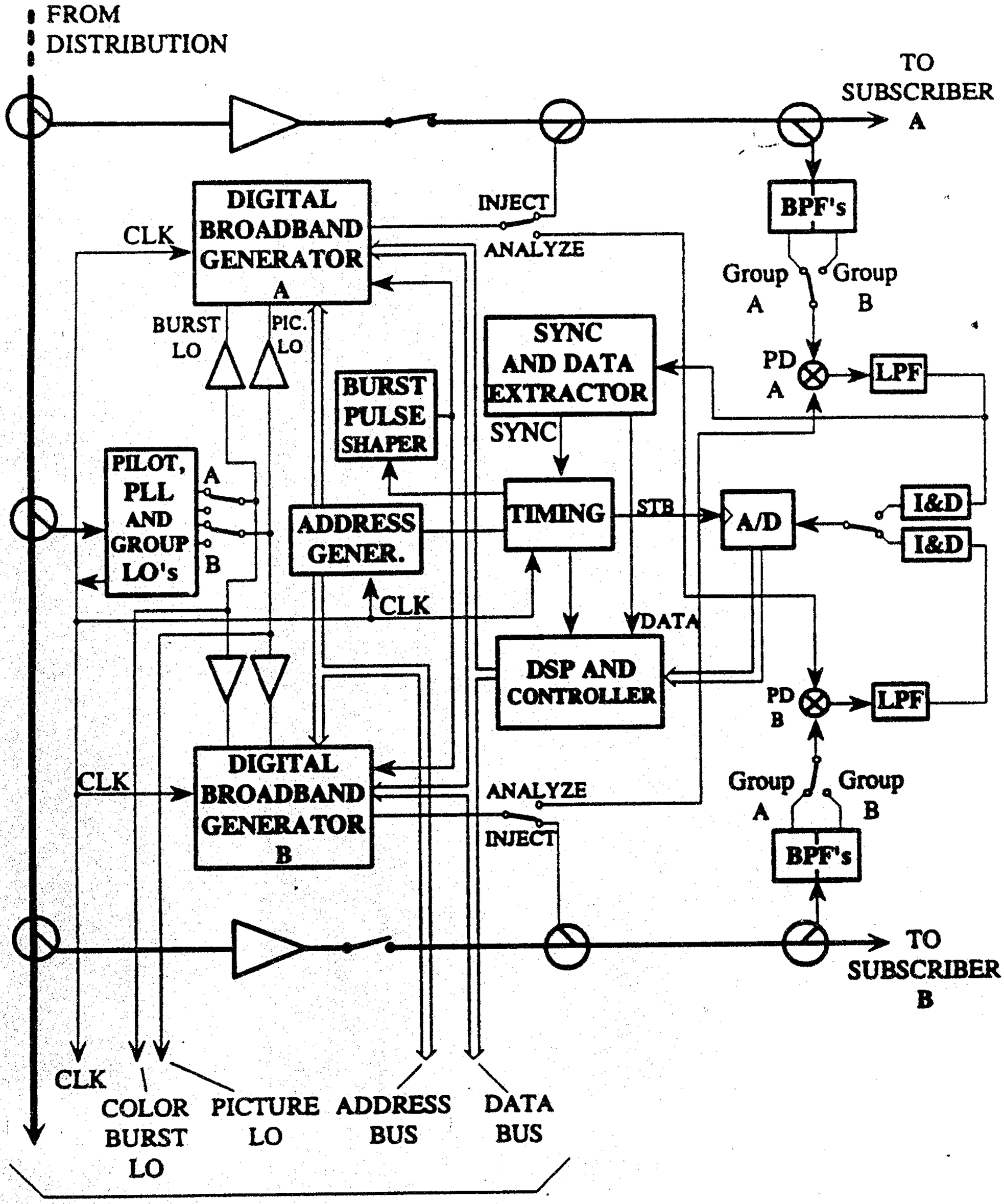

B1 RAM (PICTURE)
DAC
LPF
0°
90°
B2 RAM (PICTURE)
DAC
LPF
B1 RAM (BURST)
DAC
LPF
0°
90°
VAR. ATTEN.
BPF's
B2 RAM (BURST)
DAC
LPF
DATA
ADDR.
CLK
PICTURE LO
COLOR BURST LO
SHAPE CONTROL
OUTPUT

FIG. 17


FROM DISTRIBUTION
TO SUBSCRIBER A
DIGITAL BROADBAND GENERATOR A
INJECT
ANALYZE
CLK
BPF's
Group A
Group B
BURST LO
PIC. LO
SYNC AND DATA EXTRACTOR
PD A
LPF
BURST PULSE SHAPER
SYNC
PILOT, PLL AND GROUP LO's
A
B
ADDRESS GENER.
TIMING
STB
A/D
I&D
I&D
CLK
DATA
DSP AND CONTROLLER
PD B
LPF
DIGITAL BROADBAND GENERATOR B
CLK
ANALYZE
INJECT
Group A
Group B
BPF's
TO SUBSCRIBER B
CLK
COLOR BURST LO
PICTURE LO
ADDRESS BUS
DATA BUS
TO OTHER SUBSCRIBER UNITS

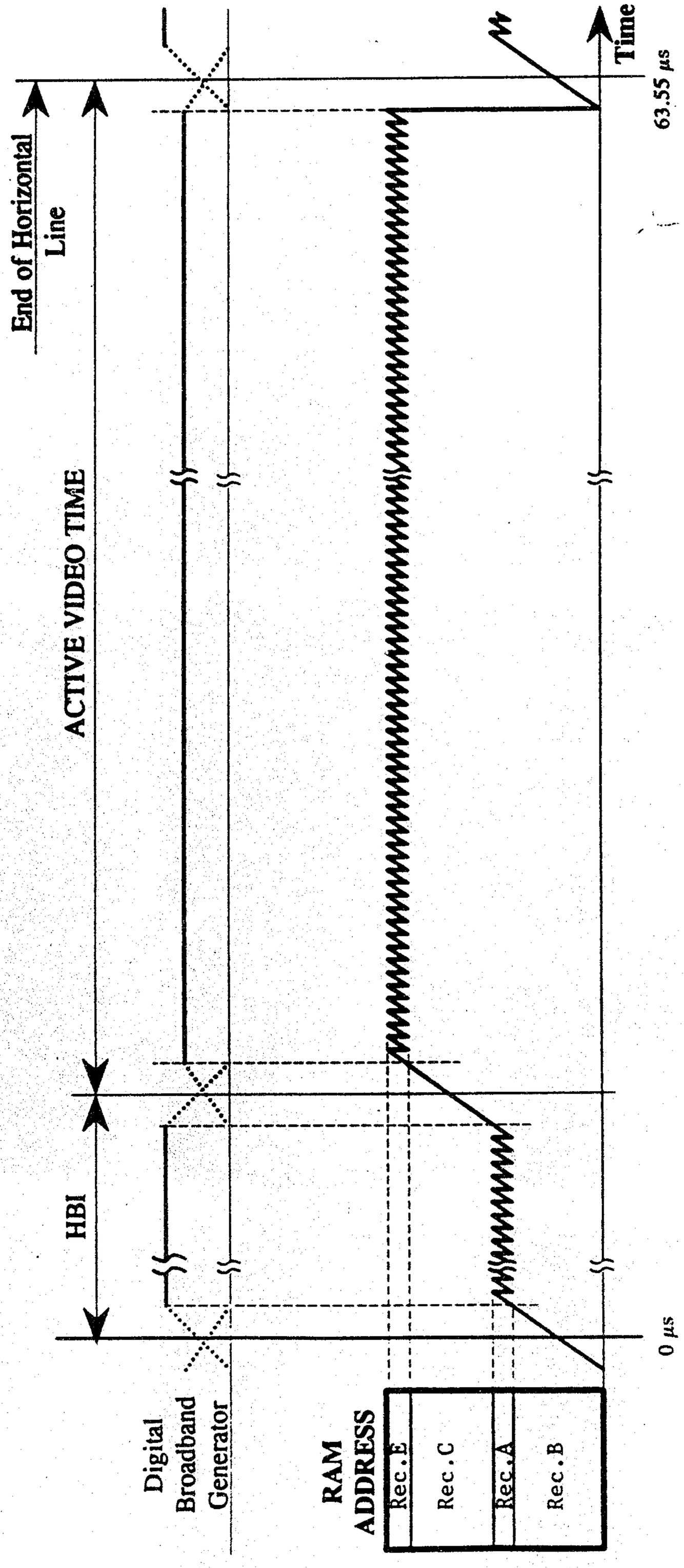
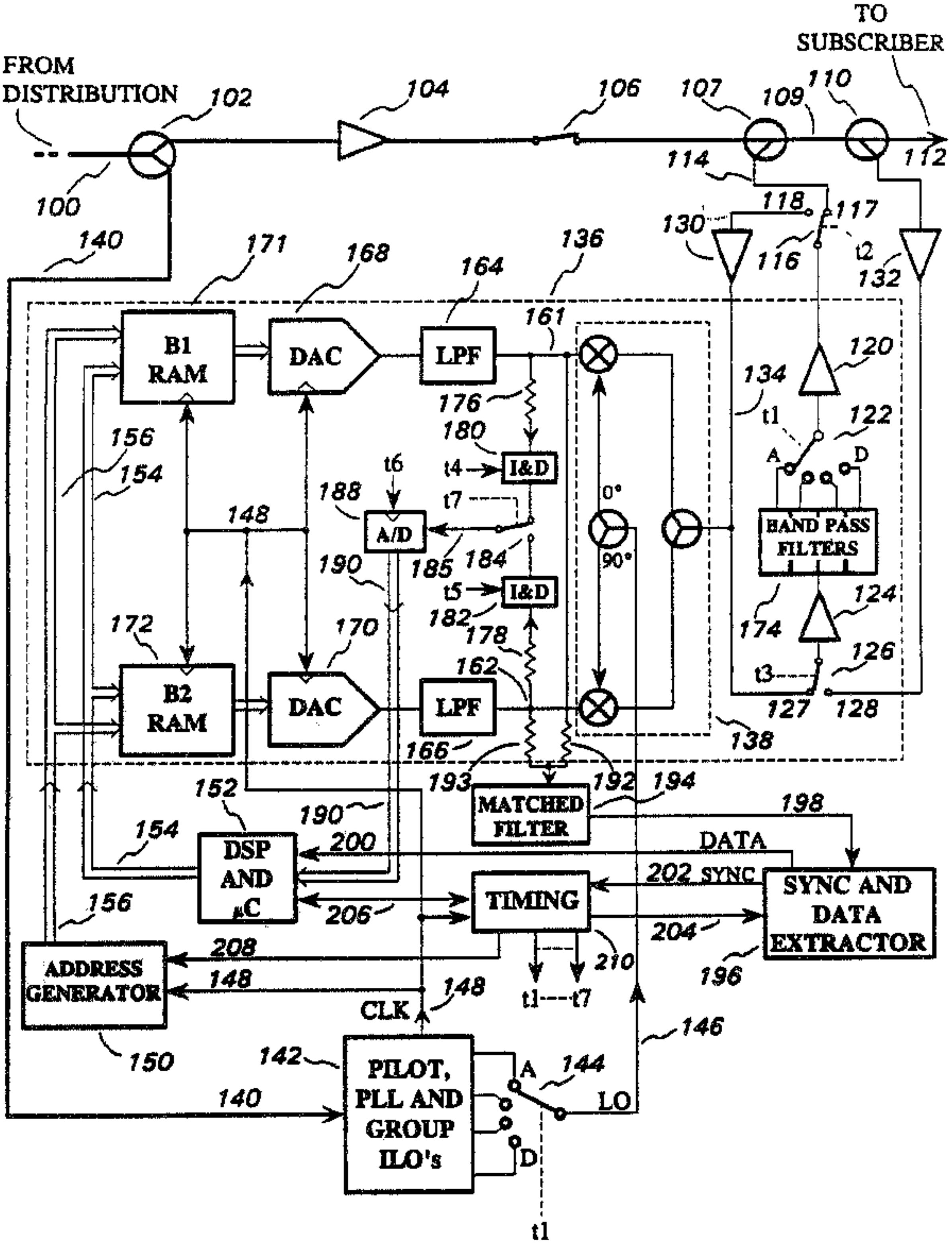

Time
63.55 µs
End of Horizontal Line
ACTIVE VIDEO TIME
HBI
0 µs
Digital Broadband Generator
RAM ADDRESS
Rec.E
Rec.C
Rec.A
Rec.B

FIG. 18


FROM DISTRIBUTION
TO SUBSCRIBER
100
102
104
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107
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110
112
114
117
118
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122
124
126
127
128
130
132
134
136
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156
161
162
164
166
168
170
171
172
174
176
178
180
182
184
185
188
190
192
193
194
196
198
200
202 SYNC
204
206
208
210
B1 RAM
B2 RAM
DAC
DAC
LPF
LPF
I&D
I&D
I&D
A/D
0°
90°
BAND PASS FILTERS
MATCHED FILTER
DSP AND µC
TIMING
SYNC AND DATA EXTRACTOR
ADDRESS GENERATOR
PILOT, PLL AND GROUP ILO's
DATA
CLK
t1---t7
t1
t3
t4
t5
t6
t7
A
D
LO