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Dependencia: 2020 DIVISION DE NUEVAS CREACIONES

001

FORMULARIO UNICO DE SOLICITUD DE PATENTE
PCT
ENTRADA EN FASE NACIONAL

① **SOLICITUD DE:**

[X] Patente de Invención [ ] Patente de Modelo de Utilidad

[ ] Capítulo I [X] Capítulo II

② **SOLICITANTE (71)**	Nombre: **ZENITH ELECTRONICS CORPORATION** Dirección: 2000 Millbrook Drive, Lincolnshire, Illinois 60069, Estados Unidos de América Nacionalidad o Domicilio: Estados Unidos de América Lugar de Constitución: Estado de Illinois, Estados Unidos de América Teléfono: 3123600 Fax: 3122410 E-mail: general@cardenasycardenas.com	**IDENTIFICACION** C.C. [ ] NIT [ ] C.E. [ ] Otro [X] Cual _____ Número _____
③ **REPRESENTANTE O APODERADO**	Nombre: **DARIO CARDENAS NAVAS** Dirección: Carrera 7 No 71-52 Torre B Piso 9 Teléfono: 3123600 Fax: 3122410 E-Mail: general@cardenasycardenas.com	**IDENTIFICACION** C.C. [X] NIT [ ] C.E. [ ] Otro [ ] Cual _____ No 17.066.629 TP 1.250
④ **INVENTOR(ES) (72)**	Nombre: THOMAS A. HORWITZ Dirección: 1236 Spring Creek Road, Elgin, Illinois 60120, Estados Unidos de América. Nacionalidad o Domicilio: Illinois, Estados Unidos de América	

⑤ **PCT (86)**

Solicitud Internacional No. PCT/US01/23211 Fecha Julio 23 de 2001
Publicación Internacional No. WO 02/13534A2 Fecha Febrero 14 de 2002

⑥ **Título(54): CODIFICADOR VSB Y MODULADOR DE RF DE BAJO COSTO PARA SUMINISTRAR UNA SEÑAL VSB SUSTANCIALMENTE DE 6 MHZ VSB A UN RECEPTOR DIGITAL DE TELEVISION.**

⑦ **Clasificación Internacional (51):** H04 N7/24.

⑧ **Prioridad** SI [X] NO [ ]

(33)País de Origen	(32)Fecha	(31)N° de Solicitud
U.S.A.	Agosto 4 de 2000	09/632.545

⑨ **Comprobante de Pago No.:** 03-4332 **Fecha** Enero 21 de 2003

10 **ANEXOS**

2

- [X] Comprobante de pago de la tasa de presentación de la solicitud.
- [X] Documento que acredita la existencia y representación legal cuando el solicitante sea persona jurídica.
- [X] Poderes, si fuere el caso.
- [ ] Documento de cesión del inventor al solicitante o a su causante. UU2
- [X] Declaraciones
- [X] Copia de la solicitud internacional. (Resumen, descripción, reivindicaciones, dibujos)
- [X] Traducción de la solicitud internacional al castellano. (Resumen, descripción, reivindicaciones, dibujos)
- [X] Copia del examen preliminar internacional efectuado por la Autoridad Internacional de Examen Preliminar (IPEA).
- [ ] Traducción del examen preliminar internacional efectuado por la IPEA.
- [ ] De ser el caso, copia del contrato de acceso.
- [ ] De ser el caso, documento que acredite la licencia o autorización de uso de conocimientos tradicionales de las comunidades indígenas.
- [ ] De ser el caso, certificado de depósito del material biológico.
- [ ] Arte final 12x12.
- [ ] De ser el caso, información sobre otras solicitudes de patente o títulos obtenidos en el extranjero por el mismo titular o su causante, relacionadas parcial o totalmente con la invención de esta solicitud.
- [ ] De ser el caso, modificaciones efectuadas a la solicitud internacional.
- [ ] Otros.

11 **FIGURA CARACTERISTICA**

12 Solicito la Concesión de la patente,

NOMBRE: DARIO CARDENAS NAVAS

FIRMA:

C.C. 17.066.629 TP. 1.250

3

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NIT : 8001760

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INCIA

RECIBO OFICIAL DE CAJA : 3 - 4,332
FECHA : ENERO 21 DE 2003

003

***** C O N S I G N A C I O N *****

DEPOSITANTE	TIPO PAGO	BANCO	CUENTA	No. PAGO	FECHA PGO	Vr. PAGO
CARDENAS Y CARDENAS P.I.	CONSIGNACION	BANCO POPULAR	050-00110-6	7972979	21/01/2003	6,648,000.00

***** C O N C E P T O *****

CANT.	RENTISTICO	CONCEPTO	TOTAL CONCEPTO
1	50005-01-01 SOLICITUDES	1 TRAMITES DE SOL. DE PATENTE DE IN	400,000.00
		TOTAL : $	400,000.00

SON: CUATROCIENTOS MIL PESOS

RESPONSABLE : _____

RECIBO DE CAJA APLICADO AL EXPEDIENTE No._____

CARDENAS & CARDENAS ABOGADOS

Zenith Electronics Corporation
*Codificador VSB y Modulador de RF de bajo costo para suministrar una señal VSB sustancialmente de 6 MHZ VSB a un receptor digital de Televisión.

Carrera 13 No. 27-00 Pisos 5, 7 y 10
Conmutador 382 0840
Fax 350 5220
E-mail: info@sic.gov.co
www.sic.gov.co
Bogotá, D.C., Colombia

- Colombia -

POWER OF ATTORNEY

In the city of Lincolnshire, on the 9th day of the month of October, of the year two thousand and two (2.002), before me, Sharon K. Kroll, Notary Public, appeared Jack Kail, of 57 age, and domiciled at Highland Park, Illinois, and stated:

FIRST: That he/she acts in his/ her capacity of Assistant Secretary of **Zenith Electronics Corporation,** a company legally constituted and in existence in accordance with the laws of the state of Illinois, and domiciled at 2000 Millbrook Drive, Lincolnshire, Illinois 60069, United States of America, as evidenced in the By-laws and Articles of Incorporation, documents which I have examined.

SECOND: That in such capacity he/she grants hereby special, ample and sufficient Power of Attorney to **DARIO CÁRDENAS NAVAS** and/or **EDUARDO CÁRDENAS CABALLERO,** and/or **BERNARDO P. CÁRDENAS MARTINEZ** and/or **JAMES VALDIRI REYES** and/or **LUZ HELENA ADARVE GÓMEZ,** and/or **JUANITA ACOSTA GÓMEZ** and/or **ANA MARÍA RUIZ RUEDA,** registered lawyers of the firm **CÁRDENAS & CÁRDENAS PROPIEDAD INTELECTUAL LAWYERS,** domiciled at Carrera 7 No. 71-52, Torre B, Piso 9 Bogotá, D.C. Colombia, to carry before the corresponding Colombian national offices and authorities the proceedings to obtain patent, industrial models and designs and trademark registrations, as well as proceedings of renewals, assignments, change of name or domicile of the owner, waivers and licenses of the same for which purpose they are empowered to take before said authorities all necessary actions for the indicated purposes; to file applications, statements, claims, formulate descriptions, amendments, oppositions and appeals; pay all taxes and installments and make any other payments determined by law; to receive every document and values issuing the respective discharge; to fulfill any other requirements and take, in general, every measure that they shall deem appropriate for the defense and protection of the trademarks of the company in the event of infringements, oppositions, proceedings for cancellation or annulment, imposition of fines, including resort to litigation; they are empowered to appear as plaintiff or defendant before the competent judges, being entitled to transact, submit to arbitration, waive, receive, resign, withdraw, appeal and all other legal powers which may be necessary, and it is hereby ratified and confirmed, any such

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INTERPRETE OFICIAL
TSR No. 823 DEL MIN. DE JUSTICIA

act said attorney lawfully do for its benefit, as my own act granting them also power to substitute this proxy and if necessary to revoke such substitutions.

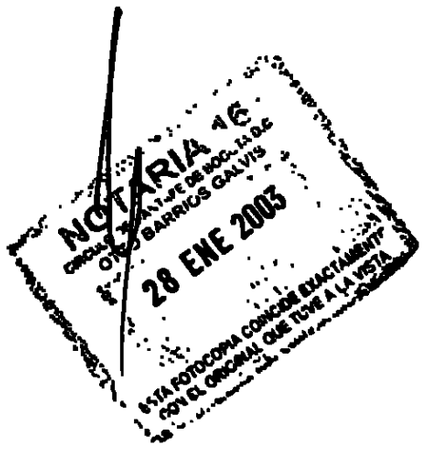
THIRD: That for purposes of articles 597 and 543 of the Colombian code of Commerce the attorneys are empowered to receive service of notice and designate judicial and extra judicial attorneys.

SIGNED AT Lincolnshire, Illinois, U.S.A.

ON October 9, 2002

BY Jack Kail
Jack Kail

Notary Public



STATE OF ILLINOIS

SECRETARY OF STATE

SPRINGFIELD, ILLINOIS

APOSTILLE

(Convention de La Haye du 5 Octobre 1961)

1. Country: United States of America

   This public document

2. has been signed by SHARON K. KROLL

3. acting in the capacity of NOTARY PUBLIC, COOK COUNTY

4. bears the seal/stamp of STATE OF ILLINOIS

Certified

5. Springfield, Illinois

6. OCTOBER 11, 2002

7. by the Secretary of State, State of Illinois

8. No. 5401

9. Seal/Stamp:

10. Signature:

Jesse White

JESSE WHITE
SECRETARY OF STATE
STATE OF ILLINOIS

NOTARIA 16 ... 28 ENE 2003 ... ESTA FOTOCOPIA CONCIDE EXACTAMENTE CON EL ORIGINAL QUE TUVE A LA VISTA

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NOTARIAL CERTIFICATION

The undersigned Notary Public certifies that Mr. Jack Kail appeared before me and signed the preceding document; that I know him personally and I am certain that he has legal capacity to grant it; that he is Assistant Secretary of **Zenith Electronics Corporation**, and is consequently authorized to grant such document; that such company is organized under the laws of the state of Illinois, and is domiciled at 2000 Millbrook Drive, Lincolnshire, Illinois 60069, United States of America; and that the act for which this document is granted is included among those which constitute the corporate object of the named company. Furthermore, that such company performs its corporate purpose in accordance with the laws of its incorporation and domicile.

I am certain of all these facts because I have examined the respective documents consisting of Colombian Power of Attorney all of which I hereby attest.

NOTARY PUBLIC (signature) Sharon K Kroll

ISSUED AND SIGNED AT Lincolnshire, Illinois

DATED October 9, 2002

OFFICIAL SEAL
SHARON K KROLL
NOTARY PUBLIC, STATE OF ILLINOIS
MY COMMISSION EXPIRES 10/30/04

NOTARIA 16
CIRCULO DE SANTAFE DE BOGOTA D.C.
OTTO BARRIOS GALVIS
28 ENE 2003
ESTA FOTOCOPIA COINCIDE EXACTAMENTE CON EL ORIGINAL QUE TUVE A LA VISTA

(LEGALIZATION BY APOSTILLE IS REQUIRED)

8

003

Traducción Oficial No. 2002/195 de un documento escrito en inglés el cual para su identificación se sella con el sello del traductor al tiempo con la presente. Esta traducción ha sido preparada por Hilda Aguilera de Piedrahita, traductora oficial debidamente reconocida por el Ministerio de Relaciones Exteriores y autorizada por el Ministerio de Justicia y del Derecho por Resolución No. 323 de 1980.

PODER

Colombia

En la ciudad de Lincolnshire, el día 9 del mes de Octubre de dos mil dos(2002), ante mí, Sharon K. Kroll Notario Público, compareció Jack Kail de 57 años de edad, y con domicilio en Highland Park, Illinois y manifestó:

PRIMERO: Que el obra en su calidad de Secretario Asistente de **Zenith Electronics Corporation,** una compañía legalmente constituida y existente de conformidad con las leyes del estado de Illinois, y con domicilio en 2000 Millbrook Drive, Lincolnshire, Illinois 60069, Estados Unidos de América, como se evidencia en los Estatutos de Constitución, documentos que he examinado.

SEGUNDO: que en tal calidad el otorga por el presente poder especial ampli y suficiente a **DARÍO CÁRDENAS-NAVAS** y/o **EDUARDO CÁRDENAS CABALLERO,** , y/o **BERNARDO P. CARDENAS MARTINEZ** y/o **JAMES VALDIRI REYES** y/o **LUZ HELENA ADARVE GOMEZ** y/o **JUANITA ACOSTA GÓMEZ** y/o **ANA MARÍA RUIZ RUEDA,** abogados registrados de la firma **CÁRDENAS & CÁRDENAS ABOGADOS PROPIEDAD INTELECTUAL LTDA.,**

NOTARIA 16 CIRCUITO DE BOGOTÁ D.C. 9 ENE 2002

ESTA FOTOCOPIA COINCIDE EXACTAMENTE CON EL ORIGINAL QUE TUVE A LA VISTA

TRADUCCION OFICIAL No. 2002/195
HILDA AGUILERA DE PIEDRAHITA
INTERPRETE OFICIAL
RESOLUCION 323 DEL MIN. DE JUSTICIA

domiciliada en la Carrera 7 No. 71-52, Torre B, Piso 9, Bogotá D.C., Colombia, para llevar a cabo ante las correspondientes Oficinas Nacionales Colombianas y autoridades los trámites para obtener patentes, modelos industriales y diseños y registros de marcas, así como trámites para renovaciones, traspasos, cambios de nombre o domicilio del propietario, renuncias y licencias de los mismos para cuyo fin están facultados para realizar ante dichas autoridades todas las acciones necesarias para los fines indicados; para presentar solicitudes, declaraciones, reivindicaciones, formular descripciones, modificaciones, oposiciones, y apelaciones; pagar todos los impuestos y derechos y hacer cualesquiera otros pagos que la ley determine; recibir todos los documentos y valores con expedición del respectivo descargo; cumplir cualesquiera otros requisitos y tomar en general todas las medidas que juzguen oportunas para la defensa y protección de las marcas de la compañía en el evento de infracciones, oposiciones, procesos de cancelación o anulación, imposición de multas, presentar antecedentes ante las cortes; quedan facultados para comparecer en calidad de demandante o demandado ante los jueces competentes, tendrán derecho de transar, someterse a arbitraje, renunciar, recibir, renunciar, desistir, apelar y todas las demás facultades legales que puedan requerirse, y por el presente declara desde ahora válido y satisfactorio lo que dichos abogados hagan en nuestro beneficio, otorgándoles también poder para sustituir este poder y si fuere necesario para revocar tales sustituciones.

NOTARIA 16
SANTAFE DE BOGOTA D.C.
BARRIOS GALAN
28 ENE 2003
ESTA FOTOCOPIA COINCIDE EXACTAMENTE CON EL ORIGINAL QUE TUVE A LA VISTA

TRADUCCION OFICIAL No. 2002/195
HILDA AGUILERA DE PEDRAZA
INTERPRETE OFICIAL
RES No. 303 DEL MINISTERIO DE HACIENDA

TERCERO: que para fines de los artículos 597 y 543 del Código Colombiano de Comercio, los abogados están facultados para recibir notificaciones y para designar apoderados judiciales y extrajudiciales.

FIRMADO EN: Lincolnshire, Illinois, EUA

EL: 9 de Octubre de 2002.

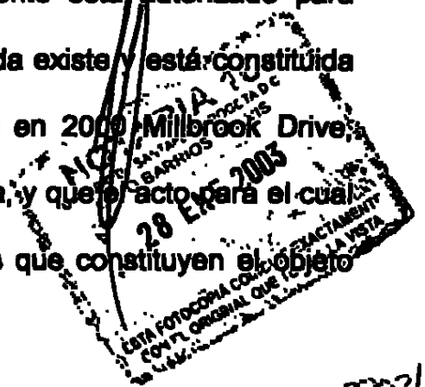
POR: Ilegible

Jack Kail

NOTARIO PUBLICO

CERTIFICACIÓN NOTARIAL

El suscrito Notario Público certifica que el Sr. Jack Kail compareció ante mí y suscribió el documento precedente; que lo conozco personalmente y que estoy seguro de que tiene las facultades para otorgarlo; que él es Secretario Asistente de **Zenith Electronics Corporation**, y por consiguiente está autorizado para otorgar dicho documento; que la compañía mencionada existe y está constituida bajo las leyes del estado de Illinois, y con domicilio en 2000 Millbrook Drive, Lincolnshire, Illinois, 60069, Estados Unidos de América, y que el acto para el cual se otorga este documento está incluido entre aquellos que constituyen el objeto social de la compañía mencionada.



2002/195

Estoy seguro de todos estos hechos por haber examinado los documentos respectivos consistentes en el Poder para Colombia, de todo lo cual doy fe.

NOTARIO PUBLICO (Firma) (Firmado:Sahron K. Kroll

EXPEDIDO Y FIRMADO EN Lincolnshire, Illinois

FECHA : Octubre 9, 2002.

Sello a tinta: SELLO OFICIAL, SHARON K. KROLL, Notario Público, Estado de Illinois,

Mi comisión expira 10/30/04.

(SE REQUIERE LEGALIZACIÓN MEDIANTE APOSTILLA)

ESTADO DE ILLINOIS

SECRETARIO DE ESTADO

SPRINGFIELD, ILLINOIS

APOSTILLA

(Convención de la Haya de Octubre 5 de 1961)

1. País: Estados Unidos de América

   Este documento público

2. Ha sido firmado por SHARON K. KROLL

3. Actuando en calidad de Notario Público, Condado de Cook

4. Lleva el sello/estampilla de l Estado de Illinois



CERTIFICADO

2002/195

5. En Springfield, Illinois

6. El día 11 de Octubre de 2002.

7. Por el Secretario de Estado del Estado de Illinois

8. Número 5401.

9. Sello (Sello Redondo en Relieve)

10. Firma (Firmado:Ilegible)

Jesse White, Secretario de Estado, Estado de Illinois.

ES TRADUCCIÓN FIEL Y COMPLETA de la cual se deja copia en los archivos de esta oficina para su confrontación y a la cual me remito.

Bogotá, Octubre 21 de 2002.

TRADUCCION OFICIAL No. 2002/195.
GILDA ABUELERA DE PEDRAHITA
INTERPRETE OFICIAL

NOTARIA 15
CIRCULO DE BOGOTA D.C.
OTTO MORRIOS GALVIS
28 ENE 2003
ESTA FOTOCOPIA CONCIDE EXACTAMENTE CON EL ORIGINAL QUE TUVE A LA VISTA

MINISTERIO DE RELACIONES EXTERIORES

023896

NO SE ASUME RESPONSABILIDAD DEL TEXTO

2002 OCT 22 P 3:00

JEFE DE LEGALIZACIONES
SANTAFE DE BOGOTA D.C.

CUYA FIRMA APARECE EN EL DOCUMENTO DESEMPEÑA LAS FUNCIONES CONSIGNADAS

NOTARIA 16
CIRCULO DE SANTAFE DE BOGOTA D.C.
OTTO BARRIOS GALVIS
28 ENE 2003
LA FOTOCOPIA CORRESPONDE EXACTAMENTE CON EL ORIGINAL QUE TUVE A LA VISTA

LOW COST VSB ENCODER AND RF MODULATOR FOR SUPPLYING A SUBSTANTIALLY 6 MHZ VSB SIGNAL TO DIGITAL TELEVISION RECEIVER

Background of the Invention and Prior Art

This invention relates generally to digital VSB (vestigial side band) television receivers and specifically to a simple low cost system for coupling a digital VSB signal from any of a plurality of sources of a digital signal (such as an MPEG - Motion Picture Experts Group - digital signal) to a digital VSB television receiver. While the description is directed to a VSB digital format, it should be understood that the invention in its broader aspects is not limited to a particular digital format.

The recently adopted Digital Television Standard specifies a VSB subsystem having a terrestrial broadcast mode (8 VSB) and a high data rate mode (16 VSB). There are other VSB modes available, namely 2 VSB, 4 VSB and 8 VSB (non terrestrial). The 8 VSB terrestrial mode has the data carrying capability of 4 VSB. (The Trellis coding that is added for the terrestrial environment creates the additional modulation levels.) The digital data signal for video is MPEG encoded and for audio is Dolby AC-3 processed and must be subjected to decompression before application to conventional video and audio circuitry. For terrestrial

broadcasting, the data signal is: randomized; subjected to Reed-Solomon (R/S) type encoding for error correction; interleaved; Trellis encoded; multiplexed with segment sync and field sync; supplied with a DC pilot; subjected to pre-equalization filtering; modulated; and RF upconverted for transmission. The digital television receiver includes a tuner and a VSB demodulator for developing the baseband signal, which is in compressed form. The demodulated signal is applied to an equalizer for equalizing the received signal. The baseband signal is applied to a transport demultiplexer which directs the data to an appropriate MPEG decoder and a Dolby decoder for recovering the video and audio in proper form for application to video and audio processing circuitry. Digital signals may be supplied to digital receivers by over-the-air transmission sources. Digital signals from other sources, such as DVD (digital video disk) players, VCR's (video cassette recorders), PC's (personal computers), digital cable boxes, satellite receivers and the like will also be supplied to digital television receivers, as is the case for present analog television receivers.

In an analog system, it is well known that coupling the signal from a VCR or other video source to a baseband input of a television receiver (if one is available) yields results superior to those achieved by modulating the signal to RF channel 3 or RF channel 4 and supplying it to the tuner input of the television receiver. This baseband coupling, however, does not necessarily produce superior results in a digital environment in which case the distance between the baseband signal source and the television receiver must be kept relatively short. In an analog system, it is also well known to supply double sideband signals to a receiver.

In accordance with the invention described in U.S. Application Serial No. 09/300,299 filed April 27, 1999, a compressed baseband digital signal is encoded in a given format for digital transmission and is modulated onto an RF carrier for direct application via a cable or wireless link to the RF input of a digital television receiver. The RF signal, which is at a low power level and has a less-than-nominal bandpass, takes advantage of the front end signal processing that is built into the digital television receiver. Signal impairments, due to the less-than-nominal

bandpass characteristic and noise introduced during transmission, that are below a given threshold are corrected by the correction circuitry in the digital television receiver front end. The low signal power, the corrective circuitry for the digital signal, and the benign environment of the communications link enable components and circuits of much lower tolerance (and cost) to be used in the encoding and modulating process implemented by the other sources described above. The result is an modulator that is very low cost and very effective in coupling a digital signal to a digital television receiver over an RF channel.

The present invention is directed to another arrangement for providing a digital signal to a digital receiver.

Summary of the Invention

In accordance with a first aspect of the invention, a system for supplying a compressed digital signal to a digital television receiver comprises a source of compressed baseband signal, an encoder, first and second frequency shifters, and first, second, and third interpolators. The encoder encodes the baseband signal in a

digital format, wherein the baseband signal has a frequency. The first frequency shifter shifts the frequency of the encoded baseband signal. The first interpolator interpolates an output of the first frequency shifter, the second interpolator interpolates an output of the first interpolator, and the third interpolator interpolates an output of the second interpolator. The second frequency shifter shifts a frequency of an output of the third interpolator.

In accordance with another aspect of the invention, a method of supplying a compressed digital signal to a digital television receiver comprises the following: encoding a baseband signal in a digital format, wherein the baseband signal has a symbol frequency; increasing the symbol frequency of the encoded baseband signal to produce an intermediate frequency signal, wherein the increasing of the symbol frequency of the encoded baseband signal includes three interpolations of the encoded baseband signal; and, modulating the intermediate frequency signal with a channel selection frequency signal so as to produce an output signal having the frequency of a selected channel.

In accordance with still another aspect of the invention, an electrical signal is produced by (i) encoding a signal in a digital television format, wherein the signal has a symbol frequency f_1, (ii) shifting the frequency of the encoded signal by $f_1/4$, (iii) increasing the frequency of the shifted encoded signal to produce an intermediate signal, and (iv) modulating the intermediate frequency signal with a channel selection frequency signal so as to produce the electrical signal at the frequency of a selected channel.

Brief Description of the Drawings

These and other features and advantages of the invention will be apparent upon reading the following description in conjunction with the drawing, in which:

Figure 1 is a simplified block diagram of a digital remodulator in accordance with the present invention;

Figure 2 is a block diagram showing the VSB digital signal modulator of Figure 1 in additional detail;

Figure 3 is a more detailed block diagram of the VSB digital signal modulator of Figure 2;

Figure 4 is a curve illustrating the frequency spectrum of the encoded baseband signal supplied to the VSB digital signal modulator of Figures 2 and 3;

Figure 5a illustrates the frequency response of the sinx/x correction filter of Figures 2 and 3 for channel 3;

Figure 5b illustrates the frequency response of the sinx/x correction filter of Figures 2 and 3 for channel 4;

Figure 6 illustrates the frequency response of the low pass filters of Figures 2 and 3;

Figure 7a illustrates the frequency response of the half band filter HBF 1 of Figures 2 and 3;

Figure 7b illustrates the frequency response of the half band filter HBF 2 of Figures 2 and 3;

Figure 7c illustrates the frequency response of the half band filter HBF 3 of Figures 2 and 3;

Figure 8a illustrates the frequency spectrum of the modulated data for channel 3;

Figure 8b illustrates the frequency spectrum of the modulated data for channel 4;

Figure 9a illustrates the frequency spectrum of the D/A output for channel 3;

Figure 9b illustrates the frequency spectrum of the D/A output for channel 4; and,

Figure 10 is a block diagram of an output circuit for the VSB digital signal modulator of Figures 2 and 3.

Description of the Preferred Embodiment

As shown in Figure 1, a source 10 provides a digital MPEG encoded signal. The source 10 may be any of a number of different types of sources. For example, the source 10 may comprise a satellite receiver, a VCR, a DVD, a digital cable box, a PC, etc. The output of the source 10 is an MPEG (and Dolby AC-3) encoded digital signal at baseband frequency. This baseband signal may be applied directly to a baseband input of a digital television receiver. However, in accordance with the present invention, the baseband signal is applied to an ATSC encoder 12 which subjects the signal to interleaving, derandomizing, R/S error encoding, and/or Trellis encoding. Field and segment syncs are added and, while not indicated, a pilot is inserted and equalization filtering is employed. (See ATSC

Standard A/53, ATSC Digital Television Standard for a detailed description of these operations.)

The ATSC encoder 12 supplies symbol data to a digital modulator 14 which remodulates the symbol data onto a selected RF channel, such as channel 3 or 4. Following remodulation, an output circuit 16 processes the remodulated data and supplies the processed remodulated data over an RF output to a digital receiver such as a digital television. The symbol data provided by the ATSC encoder 12 has a 10.76 MHZ data rate and its spectrum is flat across the entire frequency band. This data has a small DC offset which is used as a pilot, and the symbol data width is seven bits. Figure 4 shows the symbol data frequency spectrum taken with a 128 point discrete Fourier Transform (DFT).

As shown in Figures 2 and 3, a sinx/x correction filter 20 is provided at a front end of the digital modulator 14. The sinx/x correction filter 20 may have seven taps and is provided to compensate for the sinx/x roll off created by the D/A converter discussed below. Because channel 3 requires a different sinx/x correction than does channel 4, the sinx/x correction filter 20 contains two sets of tap values selected by the channel select signal. For

example, if the channel select signal is a logic level zero, the channel 3 taps are selected and, if the channel select signal is a logic level one, the channel 4 taps are selected.

The bits produced by application of these taps to the symbol data may be rounded to eleven bits and the most significant of these eleven bits may be discarded leaving a sinx/x correction output having ten bits. Even if this most significant bit were not discarded, the output of the sinx/x correction filter 20 will not exceed ten bits when the sinx/x correction filter 20 receives normal symbol data. Therefore, the most significant bit may be discarded without risk of overflow. Table 1 below shows an exemplary set of tap values for the sinx/x correction filter 20.

Table 1

Tap	Ch 3 Value	Ch 4 Value
0	-3	-4
1	1	2
2	-22	-30
3	400	400
4	-22	-30
5	1	2
6	-3	-4

The frequency response of the sinx/x correction filter 20 based upon the exemplary set of channel 3 tap values is shown in Figure 5a, and the frequency response of the sinx/x correction filter 20 based upon the exemplary set of channel 4 tap values is shown in Figure 5b.

The output of the sinx/x correction filter 20 is shifted by 2.69 MHZ, which is one-fourth of the symbol rate f_{sym}. This shift in frequency moves the DC pilot to the center of the spectrum shown in Figure 4. The frequency shift is performed by multiplying the symbol data by $e^{j1/4\omega sym}$. This multiplication is implemented in Figure 2 by a multiplier 22 which multiplies the data by $\cos(\omega_{sym}n/4)$

and a multiplier 24 which multiplies the data by $\sin(\omega_{sym}n/4)$ because

$$e^{\frac{jn\omega_{sym}}{4}} = \cos(\frac{\omega_{sym}n}{4}) + j\sin(\frac{\omega_{sym}n}{4}). \quad (1)$$

Because the frequency shift is equal to one-fourth the symbol clock rate, the factor $\cos(\omega_{sym}n/4)$ applied by the multiplier 22 produces the following sequence:

$$1,0,-1,0,1,0,-1,0 \ . \ . \ .$$

and the factor $\sin(\omega_{sym}n/4)$ applied by the multiplier 24 produces the following sequence:

$$0,1,0,-1,0,1,0,-1 \ . \ . \ . \ .$$

The two sequences can be interleaved because every other value in both sequences is zero. The resulting sequence is given by the following:

$$1,1,-1,-1,1,1,-1,-1 \ . \ . \ . \ .$$

As shown in Figure 3, this combined sequence may be implemented as a 2's complement circuit 26. The control signal from the timing block 28 turns on the 2's complement circuit 26 for two symbol clock cycles and off for two symbol clock cycles.

After the data is shifted in frequency, the real and imaginary parts of the data are filtered by low pass filters 30 and 32 as shown in Figure 2. As shown in Figure 3, the low pass filters 30 and 32 may be implemented, for example, as a single thirty-three tap low pass filter 34 because the real and imaginary data are interleaved at the output of the 2's complement circuit 26. The low pass filter 34 shapes the VSB output spectrum and has a filter roll-off that approximates the root-raised cosine roll-off of the ideal VSB output spectrum. The 3 db point of the low pass filter 34 is at 2.69 MHZ. In order to reduce the size of the low pass filter 34, the bandwidth of the low pass filter 34 is slightly wider than the 3 MHZ bandwidth of an ideal root-raised cosine filter. Table 2 below shows an exemplary set of tap values for the low pass filter 34, assuming that it is implemented as a thirty-three tap filter.

Table 2

Tap	Value	Tap	Value	Tap	Value
0	-5	11	66	22	-20
1	-14	12	22	23	-41
2	-7	13	-122	24	17
3	14	14	-24	25	27
4	11	15	384	26	-14
5	-18	16	631	27	-18
6	-14	17	384	28	11
7	27	18	-24	29	14
8	17	19	-122	30	-7
9	-41	20	22	31	-14
10	-20	21	66	32	-5

Figure 6 shows the frequency response of the low pass filter 34 based upon this exemplary set of thirty-three tap values.

As discussed previously, the real and imaginary data are interleaved at the output of the 2's complement circuit 26. When the real data are at the even taps of the low pass filter 34, the imaginary data are at the odd taps of the low pass filter 34, and vice versa. Therefore, the low pass filter 34 has two outputs, an even tap output and an odd tap output. Both outputs are interleaved with the real and imaginary data, and these outputs may be rounded to

eleven bits and limited to ten bits before being applied to a two-to-one mux 36. Such limiting will rarely occur.

In order to modulate the data, the sample rate should be increased to 86.08 MHZ, which is eight times the symbol data rate. The sample rate is increased in three stages, with each stage interpolating the data by two. Each interpolation stage includes a half band filter which removes the image of the data caused by the interpolation. Fewer filter taps are required if the interpolation and filtering are done in three steps instead of one.

Accordingly, as shown in Figure 2, the outputs of the low pass filters 30 and 32 are supplied to corresponding interpolation and half band filters 38 and 40 which perform an interpolation by two and which filter the interpolation results. The outputs of the interpolation and half band filters 38 and 40 are then supplied to corresponding interpolation and half band filters 42 and 44 which also perform an interpolation by two and which filter the interpolation results, and the outputs of the interpolation and half band filters 42 and 44 are supplied to corresponding interpolation and half band filters 46 and 48

which likewise perform an interpolation by two and filter the interpolation results.

Interpolation by two inserts a zero in between each complex data symbol. The real and imaginary parts of the data symbols are then interleaved because every other data symbol is zero. Therefore, as shown in Figure 3, only one half band filter 50 and the two-to-one mux 36 are required for the interpolation and half band filters 38 and 40, only one half band filter 52 and a two-to-one mux 56 are required for the interpolation and half band filters 42 and 44, and only one half band filter 54 and a two-to-one mux 58 are required for the interpolation and half band filters 46 and 48.

The two-to-one mux 36, under control of the timing block 28, interleaves the data by combining the even and odd tap outputs of the low pass filter 34 into one data stream with the imaginary data interleaved before the real data. The diagram below illustrates the data before and after the two-to-one mux 36.

even tap output	i_3	–	r_2		i_1	–	r_0							
	↓		↓		↓		↓	–	r_3	i_3	r_2	i_2	r_1	i_1	r_0	i_0	*mux output*
odd tap output	r_3		i_2	–	r_1		i_0							

The half band filter 50 is implemented, for example, as a filter having eleven taps. Table 3 below shows an exemplary set of tap values for the half band filter 50, assuming that it is implemented as an eleven tap filter.

Table 3

Tap	Value	Tap	Value
0	29	6	619
1	0	7	0
2	-134	8	-134
3	0	9	0
4	619	10	29
5	1024		

Figure 7a shows the frequency response for the half band filter 50 based upon this set of exemplary tap values.

The two-to-one mux 56, under control of the timing block 28, interleaves the data from the half band filter 50 by combining the even and odd tap outputs of the half band filter 50 into one data stream with the imaginary data interleaved after real data. The diagram below illustrates the data before and after the two-to-one mux 56.

$$\begin{array}{lllll} \textit{even tap output} & i_3 & r_2 - i_1 & r_0 & \\ & \vdots & \vdots & \vdots & \vdots - i_3\ r_3\ i_2\ r_2\ i_1\ r_1\ i_0\ r_0 \quad \textit{mux output} \\ \textit{odd tap output} & r_3 - i_2 & r_1 - i_0 & & \end{array}$$

The half band filter 52 is implemented, for example, as a filter having seven taps. Table 4 below shows an exemplary set of tap values for the half band filter 52, assuming that it is implemented as an seven tap filter.

Table 4

Tap	Value	Tap	Value
0	-9	4	73
1	0	5	0
2	73	6	-9
3	128		

Figure 7b shows the frequency response for the half band filter 52 based upon this set of exemplary tap values.

The two-to-one mux 58, under control of the timing block 28, interleaves the data from the half band filter 52 by combining the even and odd tap outputs of the half band filter 52 into one data stream with the imaginary data interleaved after real data. The diagram below illustrates the data before and after the two-to-one mux 58.

$$\begin{array}{ll} \textit{even tap output} & i_3 \quad r_2 - i_1 \quad r_0 \\ & \vdots \quad \vdots \quad \vdots - i_3\, r_3\, i_2\, r_2\, i_1\, r_1\, i_0\, r_0 \quad \textit{mux output} \\ \textit{odd tap output} & r_3 - i_2 \quad r_1 - i_0 \end{array}$$

The half band filter 54 is implemented, for example, as a filter having seven taps. Table 5 below shows an exemplary set of tap values for the half band filter 54, assuming that it is implemented as an seven tap filter.

Table 5

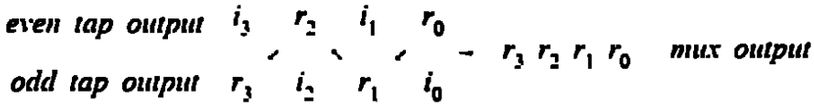
Tap	Value	Tap	Value
0	-1	4	9
1	0	5	0
2	9	6	-1
3	16		

Figure 7c shows the frequency response for the half band filter 54 based upon this set of exemplary tap values.

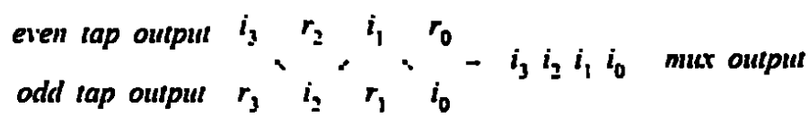
The frequency response of each of the half band filters 50, 52, and 54 is 6 db down at half of the Nyquist frequency for each of the half band filters 50, 52, and 54. Also, each of the half band filters 50, 52, and 54 has only one odd tap. Because the real and imaginary data are interleaved, the real data are in the even taps when the

imaginary data are in the odd taps and vice versa. Therefore, the half band filters 50, 52, and 54 also have two outputs, an even tap output and an odd tap output. Both outputs of each of the half band filters 50, 52, and 54 are interleaved with the real and imaginary data. The outputs of each of the two-to-one multiplexers are rounded to eleven bits and limited to ten bits.

Two two-to-one muxes 60 and 62 arranged in parallel follow the half band filter 54. The two-to-one muxes 60 and 62, under control of the timing block 28, convert the two data streams interleaved with real and imaginary data to two data streams that are either real or imaginary. The diagram below illustrates the data before and after the two-to-one mux 60.



Similarly, the diagram below illustrates the data before and after the two-to-one mux 62.



At the two-to-one mux 36, the imaginary data is inserted before the real data. This operation causes the real data and imaginary data to be skewed by one clock cycle. The two-to-one mux 56 and the two-to-one mux 58 insert the real data before the imaginary data so that the one clock cycle skew does not grow. A ten-bit register 64 shown in Figure 3 is placed after the two-to-one mux 62 in order to provide a one clock cycle delay to remove this skew.

A numerically controlled oscillator 66 has two outputs, an eight-bit sine wave output and an eight-bit cosine wave output. The phase accumulator of the numerically controlled oscillator 66 is sixteen bits. For channel 3, the phase increment is 17,582, and the output frequency is 23.098348 MHZ. For channel 4, the phase increment is 13,015, and the output frequency is 17.098453 MHZ. The frequency of the pilot (f_{out}) is determined according to the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1) \qquad (2)$$

where f_1 represents the symbol frequency, where the factor $8f_1$ results from the three interpolations-by-two, where the factor $f_1/4$ results from the 2's complement circuit 26, and where the factor f_2 represents the output of the numerically controlled oscillator 66. Accordingly, the pilot frequency for channel 3 is 60.308995 MHZ, and the pilot frequency for channel 4 is 66.308890 MHZ. Because the phase accumulator is only sixteen bits, the pilot frequencies are slightly below the ideal pilot frequencies. The channel select signal controls the NCO frequency. For example, a logic level zero selects channel 3, and a logic level one selects channel 4.

The complex data is modulated by the quadrature output of the numerically controlled oscillator 66. A multiplier 68 multiplies the ten-bit real component of the symbol data from the two-to-one mux 60 by the eight-bit cosine output of the numerically controlled oscillator 66. A multiplier 70 multiplies the ten-bit imaginary component of the symbol data from the two-to-one mux 62 and the ten-bit register 64 by the eight-bit sine output of the numerically controlled oscillator 66. A summer 72 subtracts the output of the multiplier 70 from the output of the

multiplier 68 which is equivalent to shifting the frequency spectrum of the complex data and its conjugate by the NCO frequency and summing the two together. This operation is shown by the following equation:

$$x_n = \frac{(r_n + ji_n)e^{j n\omega_{nco}} + (r_n - ji_n)e^{-jn\omega_{nco}}}{2} \quad (3)$$

$$= r_n\cos(\omega_{nco}n) - i_n\sin(\omega_{nco}n)$$

The modulated data has a center frequency equal to the output frequency of the numerically controlled oscillator 66. The pilot frequency is 2.69 MHZ above the output frequency of the numerically controlled oscillator 66. Figure 8a shows the frequency spectrum of the modulated data on channel 3, and Figure 8b shows the frequency spectrum of the modulated data on channel 4.

The output of the summer 72 is attenuated by an attenuator 74 and converted to an analog signal by a digital-to-analog converter 76. The attenuation due to the sinx/x roll-off of the digital-to-analog converter 76 can be greater for channel 4 than it is for channel 3. Also, the DC gain of the sinx/x correction filter 20 can be slightly

higher for channel 3 than for channel 4. Therefore, if these conditions exist, the signal power after the digital-to-analog converter 76 is greater (for example, by about 3 db) for channel 3 than for channel 4. Accordingly, the attenuator 74 is provided to apply a suitable attenuation (such as 23/32) to make the output power for channel 3 and channel 4 substantially equal. Thus, when the channel select signal indicates channel 3, the attenuation applied by the attenuator 74 is made active and, when the channel select signal indicates channel 4, the attenuation applied by the attenuator 74 is made inactive. The attenuator 74 may also be arranged to round the data to nine bits and to limit the data to eight bits regardless of channel.

The digital-to-analog converter 76 has an eight bit resolution and a differential output. This differential output is converted to a single ended output through a 1:1 transformer 78 of the output circuit 16 as shown in Figure 10. The side of the 1:1 transformer 78 coupled to the digital-to-analog converter 76 has a grounded center tap and the output impedance of the digital-to-analog converter 76 is set by placing a 75Ω resistor across the output of the digital-to-analog converter 76.

Figure 9a shows the frequency spectrum at the output of the digital-to-analog converter 76 for channel 3, and Figure 9b shows the frequency spectrum at the output of the digital-to-analog converter 76 for channel 4. The frequency of the numerically controlled oscillator 66 may be chosen so that the first image of the modulated data above the Nyquist frequency falls substantially in either channel 3 or channel 4. The sinx/x correction imposed by the sinx/x correction filter 20 may be also designed to correct this image. The signal-to-noise ratio of the first image is also good enough to be received by the receiver. Therefore, the first image may be used as the modulated output and the fundamental output of the digital-to-analog converter 76 can be ignored.

Accordingly, a bandpass filter 80 of the output circuit 16 as shown in Figure 10 is arranged to remove the modulated data below channels 3 and 4 and all images above channels 3 and 4. An attenuator 82 lowers the signal power to a nominal 0 dbmv.

Certain modifications and alternatives of the present invention have been discussed above. Other modifications and alternatives will occur to those practicing in

the art of the present invention. Accordingly, the description of the present invention is to be construed as illustrative only and is for the purpose of teaching those skilled in the art the best mode of carrying out the invention. The details may be varied substantially without departing from the spirit of the invention, and the exclusive use of all modifications which are within the scope of the appended claims is reserved.

WHAT IS CLAIMED IS:

1. A system for supplying a compressed digital signal to a digital television receiver comprising:

a source of compressed baseband signal;

an encoder that encodes the baseband signal in a digital format, wherein the baseband signal has a frequency;

a first frequency shifter that shifts the frequency of the encoded baseband signal;

a first interpolator that interpolates an output of the first frequency shifter;

a second interpolator that interpolates an output of the first interpolator;

a third interpolator that interpolates an output of the second interpolator; and,

a second frequency shifter that shifts a frequency of an output of the third interpolator.

2. The system of claim 1 wherein the encoder provides the baseband signal at a frequency f_1, and wherein the first frequency shifter shifts the baseband signal by $f_1/4$.

3. The system of claim 2 wherein the first interpolator interpolates by two, wherein the second interpolator interpolates by two, and wherein the third interpolator interpolates by two.

4. The system of claim 2 wherein the first, second, and third interpolators collectively interpolate by eight.

5. The system of claim 4 wherein the first and second frequency shifters and the first, second, and third interpolators effectively operate in accordance with the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the first, second, and third interpolators, where $f_1/4$ results from the first frequency shifter, and where f_2 is supplied by the second frequency shifter.

6. The system of claim 1 further comprising a low pass filter coupled between the encoder and the first interpolator, wherein the encoder provides the baseband signal at a frequency f_1, and wherein the low pass filter has a response that rolls off at about $f_1/4$.

7. The system of claim 6 wherein the first frequency shifter shifts the baseband signal by $f_1/4$.

8. The system of claim 7 wherein the first, second, and third interpolators collectively interpolate by eight.

9. The system of claim 8 wherein the first and second frequency shifters and the first, second, and third interpolators effectively operate in accordance with the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the first, second, and third interpolators, where $f_1/4$ results from the first frequency

shifter, and where f_2 is supplied by the second frequency shifter.

10. The system of claim 1 further comprising a sinx/x correction coupled between the encoder and the first interpolator, wherein the sinx/x correction provides compensation for the sinx/x roll off created by processing in the system.

11. The system of claim 10 further comprising a low pass filter coupled between the encoder and the first interpolator, wherein the encoder provides the baseband signal at a frequency f_1, and wherein the low pass filter has a response that rolls off at about $f_1/4$.

12. The system of claim 10 wherein the encoder provides the baseband signal at a frequency f_1, and wherein the first frequency shifter shifts the baseband signal by $f_1/4$.

13. The system of claim 12 wherein the first, second, and third interpolators collectively interpolate by eight.

14. The system of claim 13 wherein the first and second frequency shifters and the first, second, and third interpolators effectively operate in accordance with the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the first, second, and third interpolators, where $f_1/4$ results from the first frequency shifter, and where f_2 is supplied by the second frequency shifter.

15. The system of claim 1 wherein the first interpolator interpolates by two, wherein the second interpolator interpolates by two, and wherein the third interpolator interpolates by two.

16. A method of supplying a compressed digital signal to a digital television receiver comprising:

encoding a baseband signal in a digital format, wherein the baseband signal has a symbol frequency;

increasing the symbol frequency of the encoded baseband signal to produce an intermediate frequency signal, wherein the increasing of the symbol frequency of the encoded baseband signal includes three interpolations of the encoded baseband signal; and,

modulating the intermediate frequency signal with a channel selection frequency signal so as to produce an output signal having the frequency of a selected channel.

17. The method of claim 16 wherein the baseband signal has a symbol frequency f_1, and wherein the increasing of the symbol frequency of the encoded baseband signal comprises shifting the symbol frequency of the baseband signal by $f_1/4$.

18. The method of claim 16 wherein the three interpolations are each an interpolation by two.

19. The method of claim 16 wherein the baseband signal has a symbol frequency f_1, and wherein the output signal has a frequency f_{out} given by the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the three interpolations, where $f_1/4$ results from a frequency shifting of the encoded baseband signal, and where f_2 is the frequency of the channel selection frequency signal.

20. The method of claim 16 further comprising low pass filtering the encoded baseband signal, wherein the baseband signal has a symbol frequency f_1, and wherein the low pass filtering has a response that rolls off at about $f_1/4$.

21. The method of claim 20 wherein the increasing of the symbol frequency of the encoded baseband signal comprises shifting the symbol frequency of the baseband signal by $f_1/4$.

22. The method of claim 20 wherein the three interpolations are each an interpolation by two.

23. The method of claim 20 wherein the output signal has a frequency f_{out} given by the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the three interpolations, where $f_1/4$ results from a frequency shifting of the encoded baseband signal, and where f_2 is the frequency of the channel selection frequency signal.

24. The method of claim 16 further comprising sinx/x correcting the encoded baseband signal, wherein the sinx/x correction compensates for the sinx/x roll off created by processing downstream of the sinx/x correction.

25. The method of claim 24 further comprising low pass filtering the encoded baseband signal, wherein the baseband signal has a symbol frequency f_1, and wherein the low pass filtering has a response that rolls off at about $f_1/4$.

26. The method of claim 24 wherein the baseband signal has a symbol frequency f_1, and wherein the increasing of the symbol frequency of the encoded baseband signal comprises shifting the symbol frequency of the baseband signal by $f_1/4$.

27. The method of claim 24 wherein the three interpolations are each an interpolation by two.

28. The method of claim 24 wherein the baseband signal has a symbol frequency f_1, and wherein the output signal has a frequency f_{out} given by the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the three interpolations, where $f_1/4$ results from a frequency shifting of the encoded baseband signal, and where f_2 is the frequency of the channel selection frequency signal.

29. The method of claim 16 wherein the output signal contains modulated data below the selected channel and an image at the selected channel, and wherein the method comprises filtering of the output signal to remove the modulated data and to pass the image.

30. An electrical signal produced by (i) encoding a signal in a digital television format, wherein the signal has a symbol frequency f_1, (ii) shifting the frequency of the encoded signal by $f_1/4$, (iii) increasing the frequency of the shifted encoded signal to produce an intermediate signal, and (iv) modulating the intermediate frequency signal with a channel selection frequency signal so as to produce the electrical signal at the frequency of a selected channel.

31. The electrical signal of claim 30 wherein the frequency of the shifted encoded signal is increased by interpolation, and wherein the interpolation comprises three interpolations by two.

32. The electrical signal of claim 30 having a frequency f_{out} given by the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the increasing of the frequency of the shifted encoded signal, where $f_1/4$ results from the frequency shifting, and where f_2 is the frequency of the channel selection frequency signal.

33. The electrical signal of claim 30 wherein the encoded signal is low pass filtered, and wherein the low pass filtering has a response that rolls off at about $f_1/4$.

34. The electrical signal of claim 33 wherein the frequency of the shifted encoded signal is increased by interpolation, and wherein the interpolation comprises three interpolations by two.

35. The electrical signal of claim 33 having a frequency f_{out} given by the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the increasing of the frequency of the shifted encoded signal, where $f_1/4$ results from the frequency shifting, and where f_2 is the frequency of the channel selection frequency signal.

36. The electrical signal of claim 30 wherein the encoded signal is sinx/x corrected, and wherein the sinx/x correction compensates for the sinx/x roll off created by processing of the electrical signal downstream of the sinx/x correction.

37. The electrical signal of claim 36 wherein the encoded signal is low pass filtered, and wherein the low pass filtering has a response that rolls off at about $f_1/4$.

38. The electrical signal of claim 36 wherein the frequency of the shifted encoded signal is increased by interpolation, and wherein the interpolation comprises three interpolations by two.

39. The electrical signal of claim 36 having a frequency f_{out} given by the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the increasing of the frequency of the shifted encoded signal, where $f_1/4$ results from the frequency shifting, and where f_2 is the frequency of the channel selection frequency signal.

40. The electrical signal of claim 30 further produced by filtering so as remove data below the selected channel and so as to leave an image of the data as the electrical signal.

ABSTRACT OF THE DISCLOSURE

A system supplies a compressed digital signal to a digital television receiver and includes a source of compressed baseband signal. An encoder encodes the baseband signal in a digital format, and the baseband signal has a frequency. The frequency of the encoded baseband signal is shifted by a first frequency shifter. A first interpolator interpolates an output of the first frequency shifter, a second interpolator that interpolates an output of the first interpolator, and a third interpolator that interpolates an output of the second interpolator. A second frequency shifter shifts a frequency of an output of the third interpolator to produce a signal at a selected channel.

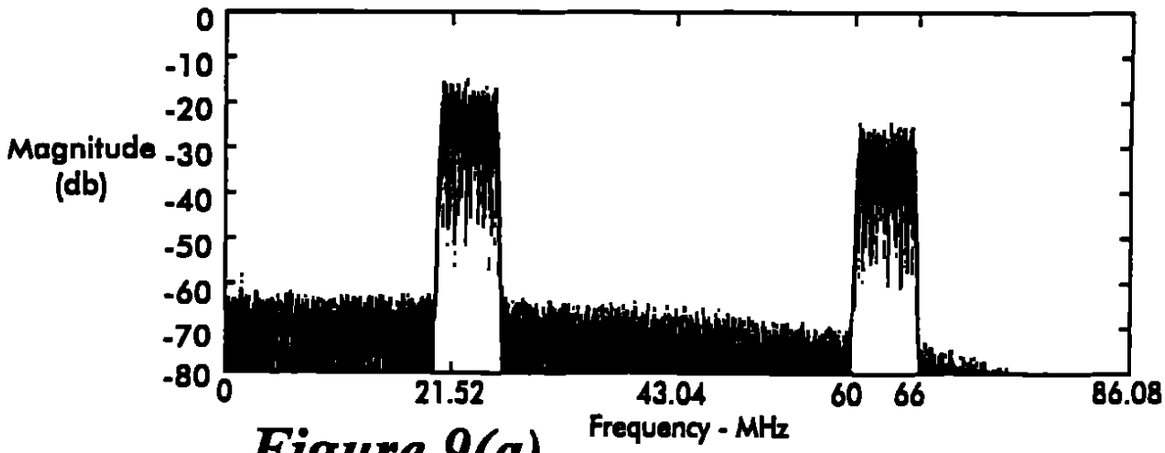


Figure 9(a)

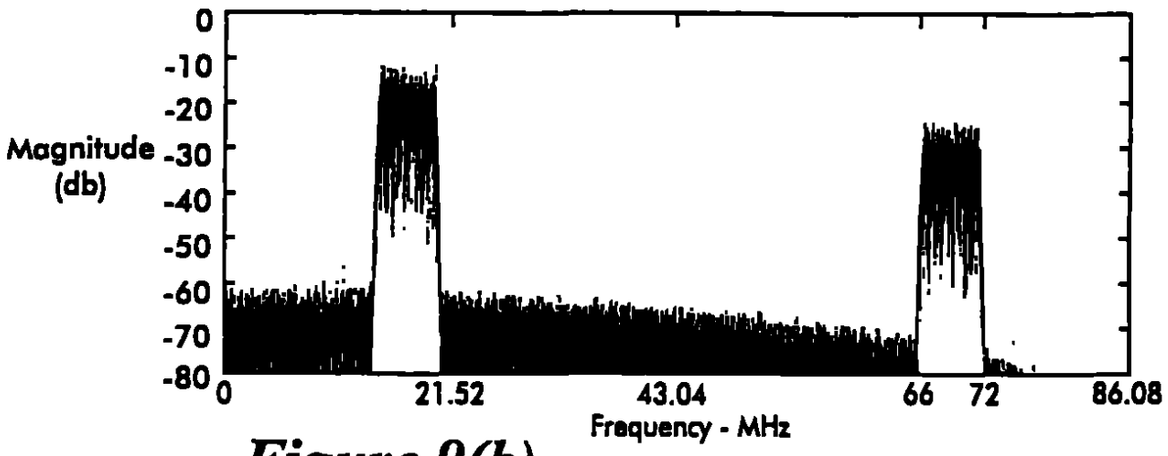


Figure 9(b)

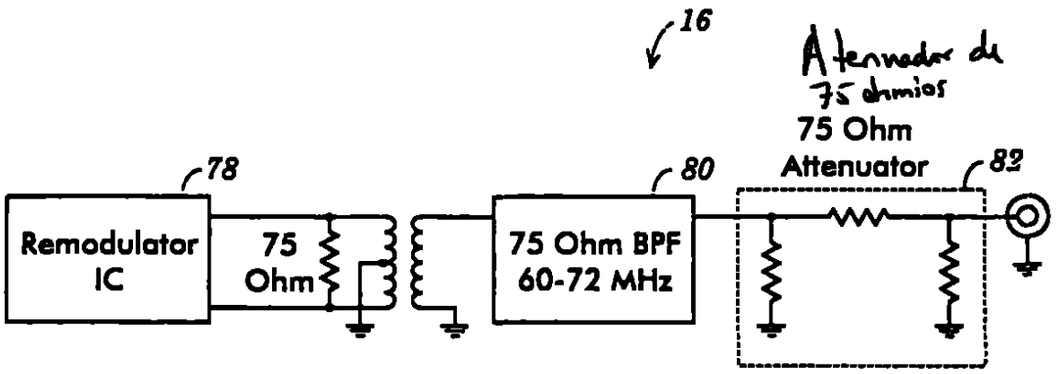


Figure 10

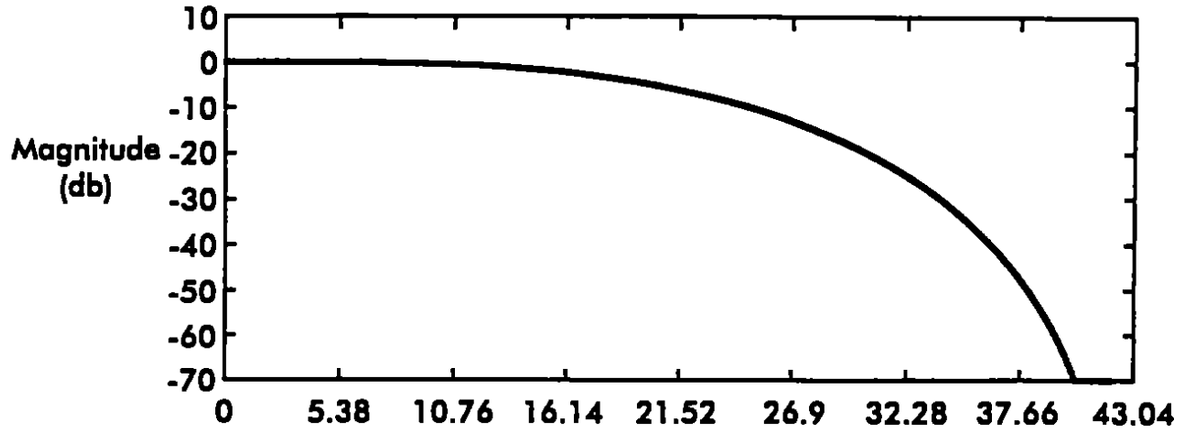


Figure 7(c)

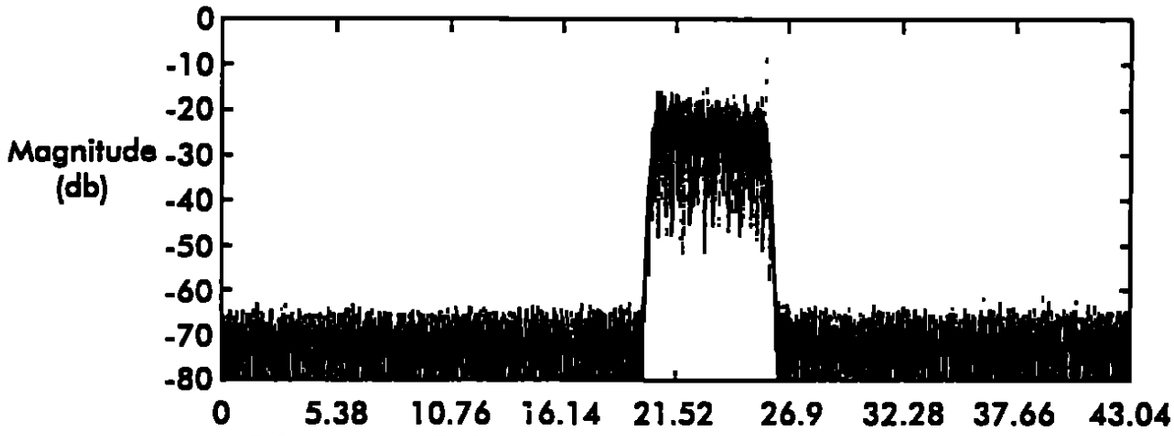


Figure 8(a)

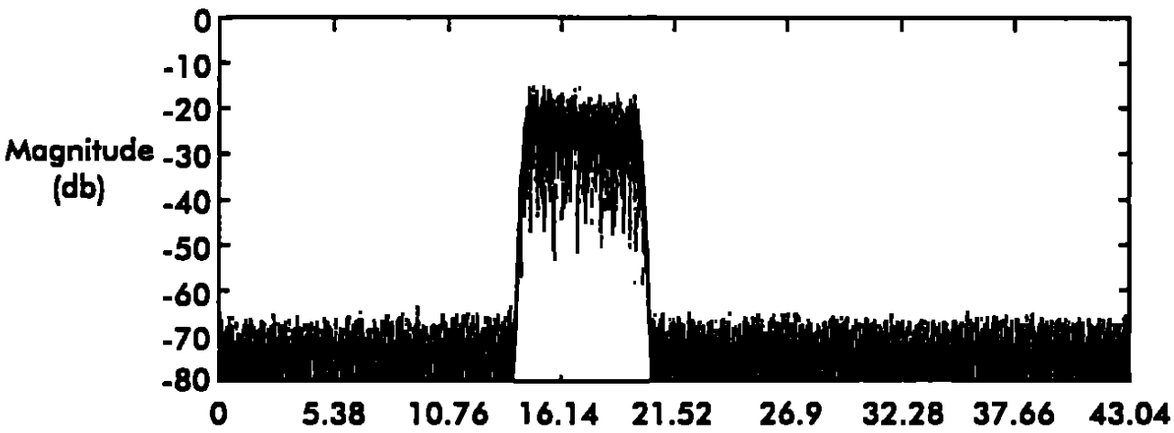


Figure 8(b)

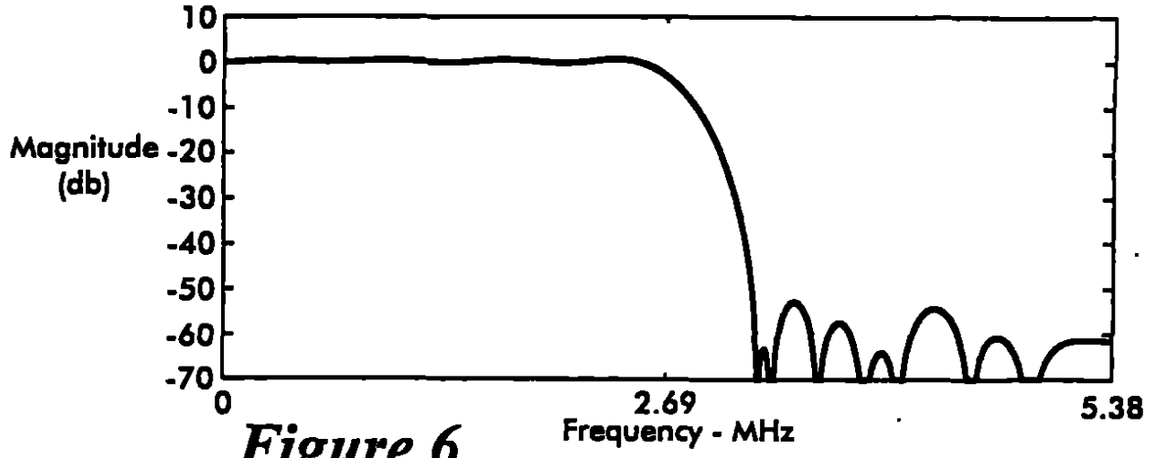


Figure 6

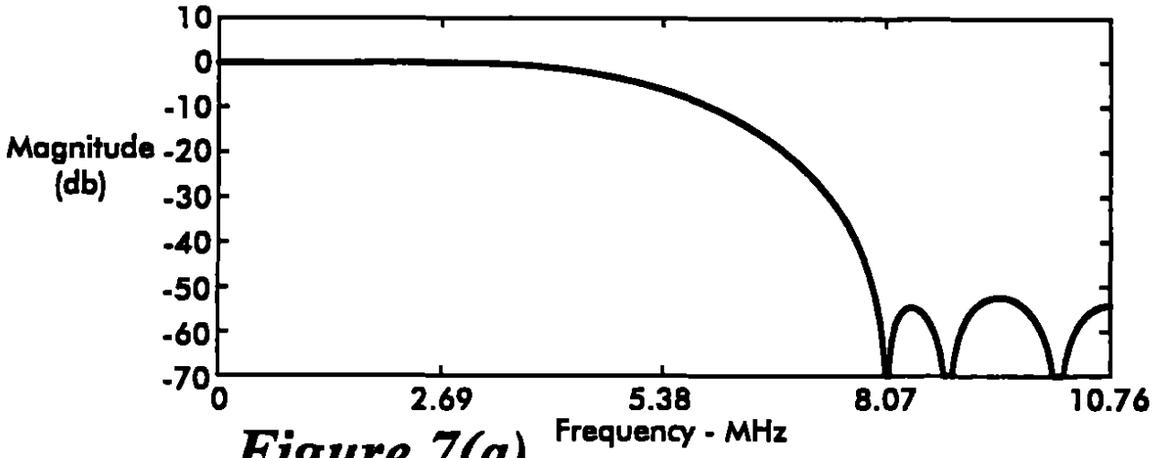


Figure 7(a)

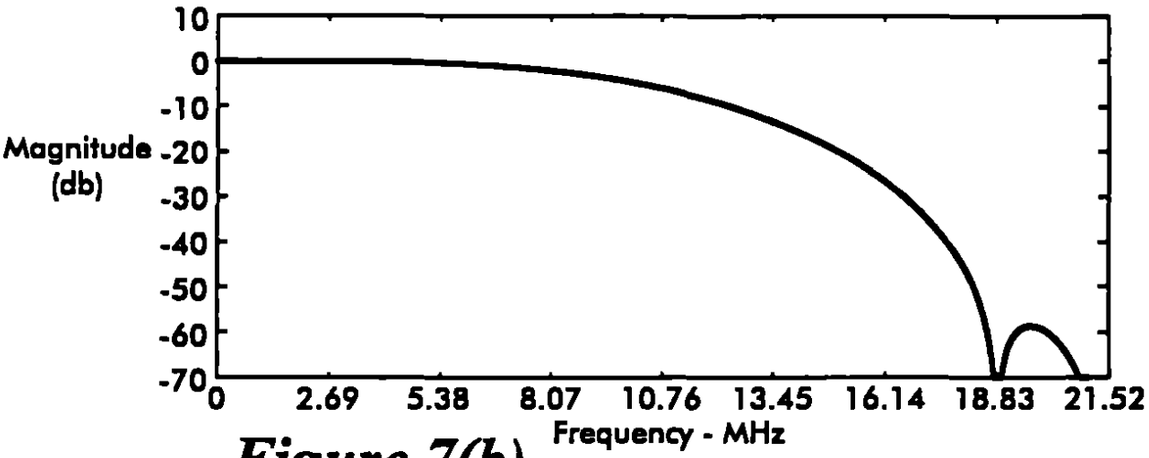


Figure 7(b)

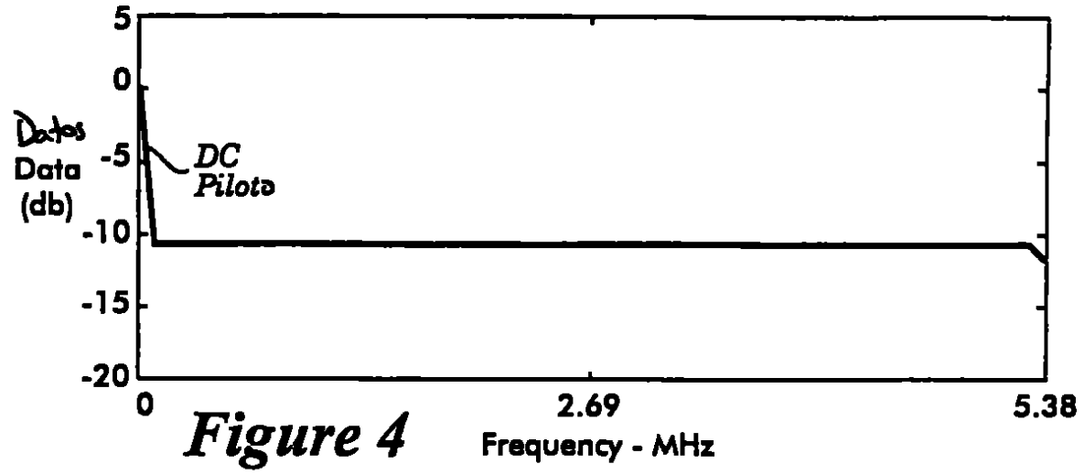


Figure 4

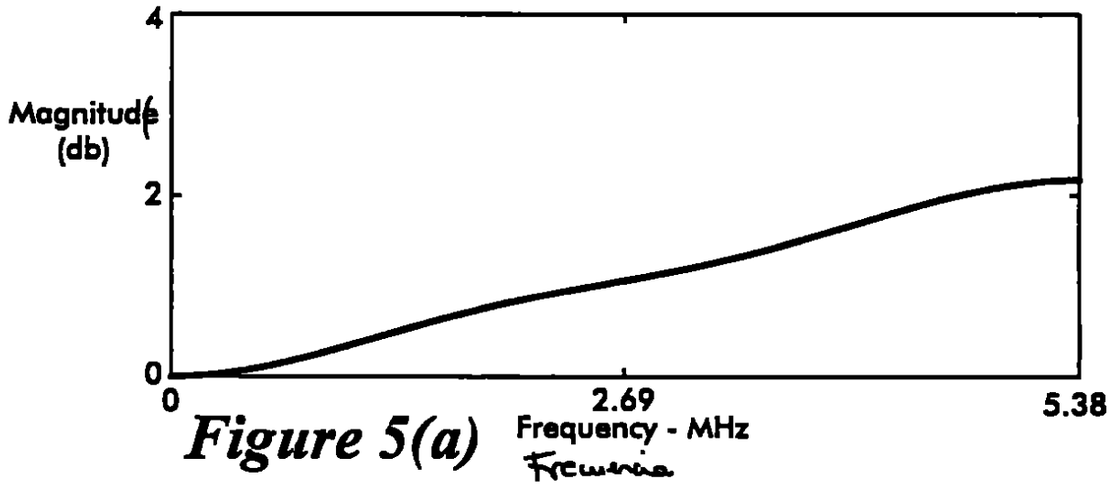


Figure 5(a)

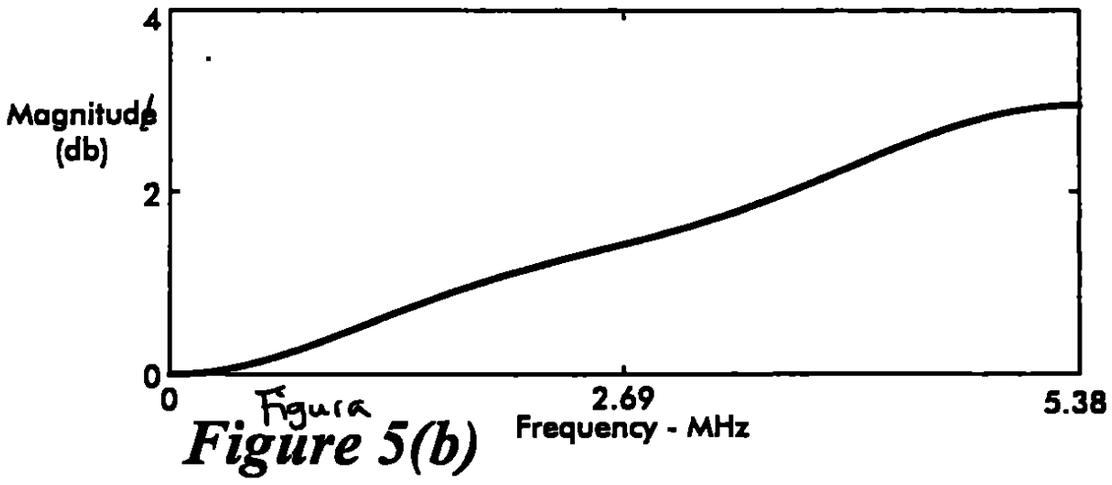


Figure 5(b)

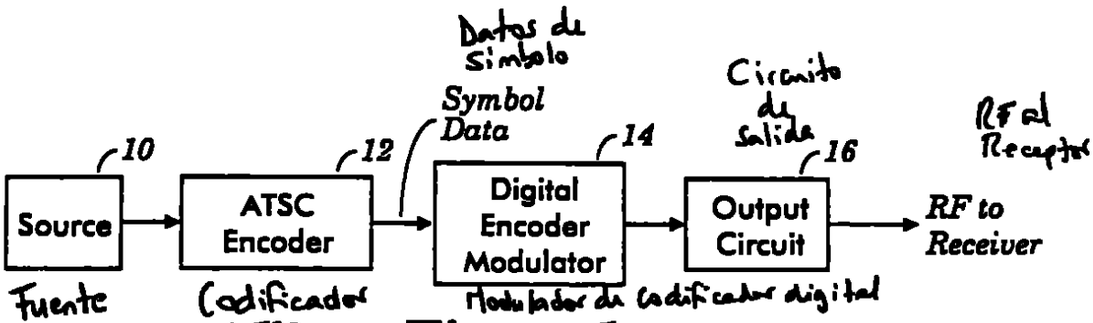


Figure 1

Figura 1

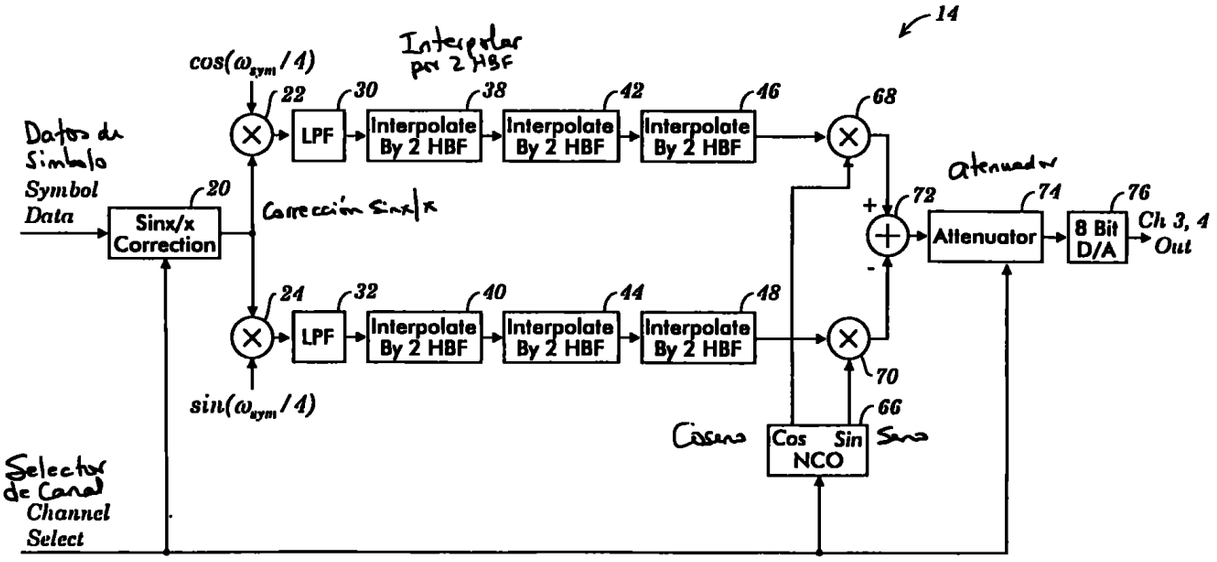
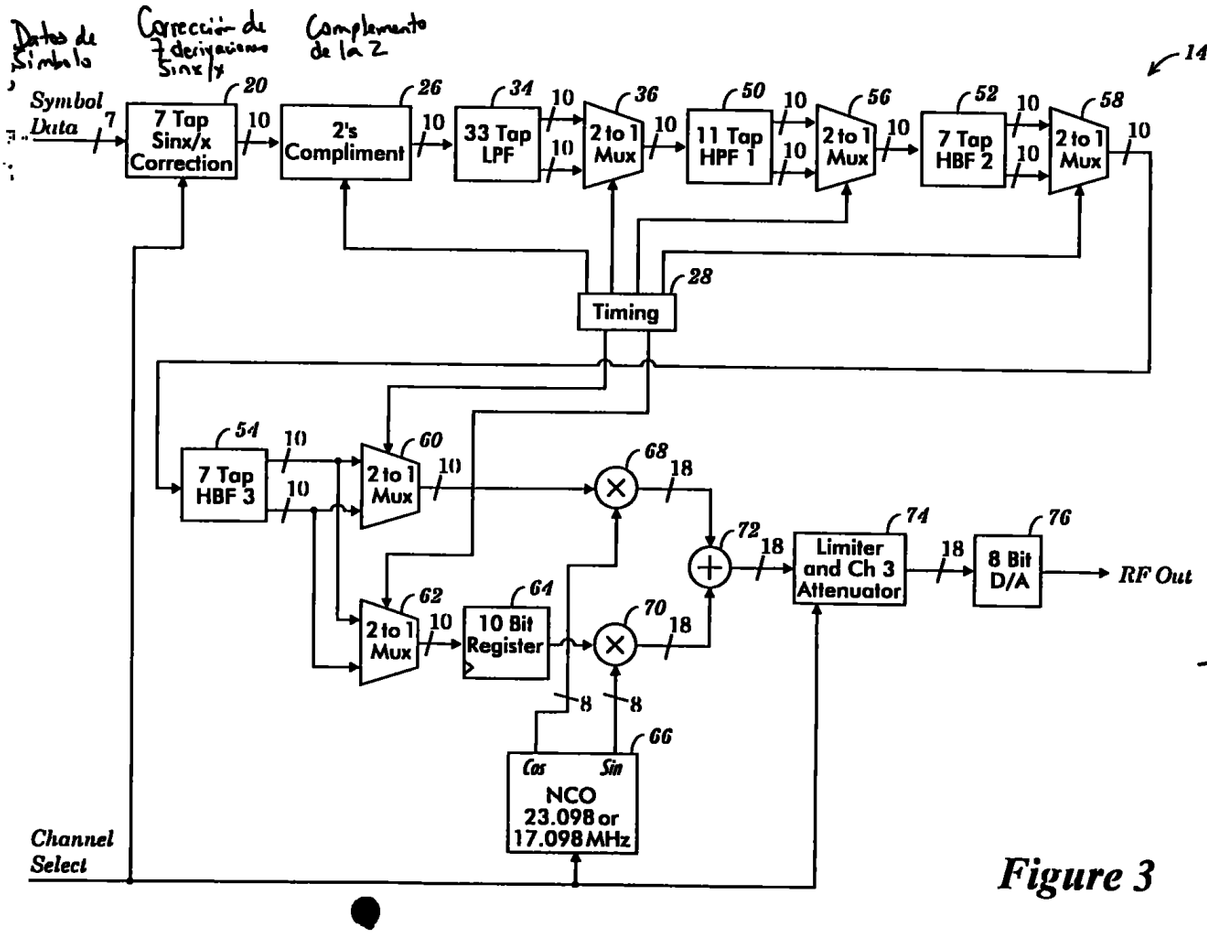


Figure 2
Figura 2



tap - derivación
mux - multiplexor

Figure 3

CODIFICADOR VSB Y MODULADOR DE RF DE BAJO COSTO PARA SUMINISTRAR UNA SEÑAL VSB SUSTANCIALMENTE DE 6 MHZ VSB A UN RECEPTOR DIGITAL DE TELEVISIÓN

Antecedentes del Invento y Técnica Anterior

Este invento se relaciona en general con receptores digitales VSB de televisión (banda lateral residual) y específicamente con un sistema de bajo costo para acoplar una señal VSB digital de cualesquiera de una pluralidad de fuentes de una señal digital (tales como una señal digital MPEG -Grupo de Expertos de Películas de Cine-) a un receptor digital VSB de televisión. Mientras que la descripción se dirige a un formato VSB digital, debe entenderse que el invento en su aspecto más amplio no se limita a un formato digital particular.

La Norma de Televisión Digital adoptada recientemente especifica un subsistema VSB que tiene un modo de difusión terrestre (8 VSB) y un modo de alta velocidad de datos (16 VSB). Existen otros modos VSB disponibles, a saber: 2 VSB, 4 VSB y 8 VSB (no terrestres). El modo terrestre 8 VSB tiene la capacidad de portar datos del 4 VSB. (La codificación de cuadrículas (Trellis) que se añade para el ambiente terrestre crea los niveles de modulación adicional). La señal de datos

digitales para video se codifica en MPEG y para el audio se procesa en Dolby AC-3 y se debe someter a descompresión antes de aplicarse al video convencional o al conjunto de circuitos de audio. Para la difusión terrestre la señal de datos es: aleatorea; sujeta a codificación tipo Reed-Solomon (R/S) para corrección de errores; intercalada; codificada en cuadrícula (Trellis); multiplexada con sincronización de segmentos y sincronización de campo; suministrada con un piloto DC; sujeta a un filtrado de pre-ecualización; modulada; y convertida a RF para transmisión. El receptor digital de televisión incluye un sintonizador y un desmodulador VSB para desarrollar la señal de la banda de base, la cual está en forma comprimida. La señal desmodulada se aplica a un ecualizador para ecualizar la señal recibida. La señal de la banda de base se aplica a un desmultiplexor de transporte que dirige los datos a un decodificador MPEG apropiado y a un decodificador Dolby para recuperar el video y el audio en forma idónea para aplicación al conjunto de circuitos de procesamiento de audio y video. Las señales digitales se pueden suministrar a receptores digitales por fuentes de transmisión entre puntos aéreos. Las señales digitales de otras fuentes, tales como tocadiscos DVD (disco de video digital), VCRs (grabadoras de videocasete), PCs (computadoras

personales), caja de cable digital, receptores de satélites y similares también se suministrarán a receptores digitales de televisión, tal como es el caso de los receptores actuales análogos para televisión.

En un sistema análogo, es bien conocido que acoplar la señal de un VCR o de otra fuente de video a una entrada de banda de base de un receptor de televisión (si se dispone de él) produce resultados superiores a aquellos logrados mediante la modulación de la señal a un canal de RF 3 o a un canal de RF 4 suministrándola a la entrada del sintonizador del receptor de televisión. Sin embargo, este acople de la banda de base no necesariamente produce resultados superiores en un ambiente digital, en cuyo caso la distancia entre la fuente de la señal de la banda de base y el receptor de televisión debe mantenerse relativamente corta. En un sistema análogo, igualmente es bien conocido suministrar señales dobles de banda lateral a un receptor.

De conformidad con el invento descrito en la Solicitud de Patente Norteamericana Serie No. 09/300.299 presentada el 27 de abril de 1999, una señal digital comprimida de banda de base se codifica en un formato dado para transmisión digital y se modula a un portador RF para aplicación directa a través

de un cable o enlace inalámbrico a la entrada de RF de un receptor digital de televisión. La señal RF, que está en un nivel de baja potencia y tiene un paso de banda inferior al nominal, aprovecha el procesamiento de la señal del extremo frontal que está incluido en el receptor digital de televisión. Los deterioros de la señal debidos a la característica del paso de banda inferior al nominal y al ruido introducido durante la transmisión, que están por debajo de un umbral dado, se corrigen por un conjunto de circuitos de corrección en el extremo frontal del receptor digital de televisión. La potencia de baja señal, el conjunto de circuitos correctivo para la señal digital, y el ambiente benigno del enlace de las comunicaciones permiten que se utilicen componentes y circuitos de tolerancia (y costos) mucho más bajos en el proceso de codificación y modulación implementado por las otras fuentes arriba descritas. El resultado es un modulador que es de muy bajo costo y muy eficaz para acoplar una señal digital a un receptor digital de televisión en un canal de RF.

El presente invento está dirigido a otro arreglo para proporcionar una señal digital a un receptor digital.

Resumen del Invento

De conformidad con un primer aspecto del invento, un sistema para suministrar una señal digital comprimida a un receptor digital de televisión comprende una fuente de señal de banda de base comprimida, un codificador, un primero y segundo desplazadores de frecuencia, y un primero, segundo y tercer interpoladores. El codificador codifica la señal de la banda de base en un formato digital, en donde la señal de la banda de base tiene una frecuencia. El primer desplazador de frecuencia cambia la frecuencia de la señal de la banda de base codificada. El primer interpolador interpola una potencia de salida del primer desplazador de frecuencia, el segundo interpolador interpola una potencia de salida del primer interpolador, y el tercer interpolador interpola una potencia de salida del segundo interpolador. El segundo desplazador de frecuencia cambia una frecuencia de una potencia de salida del tercer interpolador.

De acuerdo con otro aspecto del invento, un método de suministrar una señal digital comprimida a un receptor digital de televisión comprende lo siguiente: codificar una señal de banda de base en un formato digital, en donde la señal de banda de base tiene una frecuencia con símbolo; aumentando la frecuencia de símbolo de la señal de banda de

base codificada para producir una señal de frecuencia intermedia, en donde el aumento de la frecuencia de símbolo de la señal de banda de base codificada incluye tres interpolaciones de la señal de banda de base codificada; y, modulando la señal de la frecuencia intermedia con una señal de frecuencia de selección de canal de manera que se produzca una señal de salida que tenga la frecuencia de un canal seleccionado.

De conformidad con otro aspecto del invento, una señal eléctrica es producida por (i) la codificación de una señal en un formato digital de televisión, en donde la señal tiene una frecuencia de símbolo f_1, (ii) el desplazamiento de frecuencia de la señal codificada por $f_1/4$, (iii) el aumento de la frecuencia de la señal codificada desplazada para producir una señal intermedia, y (iv) la modulación de la señal de frecuencia intermedia con una señal de frecuencia de selección de canal de manera que se produzca una señal eléctrica en la frecuencia de un canal seleccionado.

Breve Descripción de los Dibujos

Éstas y otras características y ventajas del invento serán evidentes cuando se lea la siguiente descripción en conjunto con los dibujos, en donde:

La Figura 1 es un diagrama de bloque simplificado de un remodulador digital de conformidad con el presente invento;

La Figura 2 es un diagrama de bloque que muestra el modulador de la señal digital VSB de la Figura 1 con detalles adicionales;

La Figura 3 es un diagrama de bloque más detallado del modulador de señal digital VSB de la Figura 2;

La Figura 4 es una curva que ilustra el espectro de frecuencia de la señal de la banda de base codificada suministrada al modulador de señal digital VSB de las Figuras 2 y 3;

La Figura 5a ilustra la respuesta de la frecuencia del filtro de corrección *sinx/x* de las Figuras 2 y 3 para el canal 3;

La Figura 5b ilustra la respuesta de la frecuencia del filtro de corrección sinx/x de las Figuras 2 y 3 para el canal 4;

La Figura 6 ilustra la respuesta de la frecuencia de los filtros de paso bajo de las Figuras 2 y 3;

La Figura 7a ilustra la respuesta de la frecuencia del filtro de media banda HBF 1 de las Figuras 2 y 3;

La Figura 7b ilustra la respuesta de la frecuencia del filtro de media banda HBF 2 de las Figuras 2 y 3;

La Figura 7c ilustra la respuesta de la frecuencia del filtro de media banda HBF 3 de las Figuras 2 y 3;

La Figura 8a ilustra el espectro de la frecuencia de los datos modulados para el canal 3;

La Figura 8b ilustra el espectro de la frecuencia de los datos modulados para el canal 4;

La Figura 9a ilustra el espectro de la frecuencia de la potencia de salida D/A para el canal 3;

La Figura 9b ilustra el espectro de la frecuencia de la potencia de salida D/A para el canal 4; y

La Figura 10 es un diagrama de bloque de un circuito de salida para el modulador de señal digital VSB de las Figuras 2 y 3.

Descripción de la Incorporación Preferida

Tal como se muestra en la Figura 1, una fuente (10) proporciona una señal digital codificada MPEG. La fuente (10)

puede ser cualquiera de un número de diferentes tipos de fuentes. Por ejemplo, la fuente (10) puede comprender un receptor de satélite, un VCR, un DVD, una caja de cable digital, un PC, etc. La potencia de salida de la fuente (10) es una señal digital codificada MPEG (y Dolby AC-3) en la frecuencia de la banda de base. Esta señal de la banda de base se puede aplicar directamente a una entrada de la banda de base de un receptor digital de televisión. Sin embargo, de conformidad con el presente invento, la señal de la banda de base se aplica a un codificador ATSC (12) que somete la señal a intercalación, aleatorización, codificación de error R/S, y/o codificación de cuadrículas (Trellis). Se añaden las sincronizaciones de segmentos y de campo y, aunque no se indica, se inserta un piloto y se emplea un filtrado para ecualización. (Ver la Norma ATSC A/53, la Norma de Televisión Digital ATS para una descripción detallada de estas operaciones).

El codificador ATSC (12) suministra datos del símbolo a un modulador digital (14) que remodula los datos del símbolo hacia un canal de RF seleccionado, tal como el canal 3 o 4. Después de la remodulación, un circuito de salida (16) procesa los datos remodulados y suministra los datos remodulados procesados a través de una salida de RF a un

receptor digital, tal como una televisión digital. Los datos de los símbolos provistos por el codificador ATSC (12) tienen una velocidad de datos MHZ de 10.76 y su espectro es plano a lo largo de la totalidad de la banda de frecuencia. Estos datos tienen una pequeña compensación DC que se utiliza como piloto, y el ancho de los datos del símbolo es de 7 bits. La Figura 4 muestra el espectro de la frecuencia de los datos de símbolo tomado con un Transformador Fourier discreto (DFT, por sus iniciales en inglés) de 128 puntos.

Tal como se muestra en las Figuras 2 y 3, un filtro de corrección sinx/x (20) se suministra en el extremo frontal del modulador digital (14). El filtro de corrección sinx/x (20) puede tener siete derivaciones y se proporciona para compensar la atenuación progresiva del sinx/x creado por el convertidor D/A discutido a continuación. Como el canal (3) requiere una corrección sinx/x diferente de aquella del canal (4), el filtro de corrección sinx/x (20) contiene dos juegos de valores de derivaciones seleccionados por la señal que selecciona el canal. Por ejemplo, si la señal de selección del canal es un lógica de nivel cero, se seleccionan las derivaciones (3) del canal y, si la señal de selección de canal es un lógica de nivel uno, se seleccionan las derivaciones (4) del canal.

Los bits producidos por la aplicación de estas derivaciones a los datos del símbolo se pueden redondear a once bits y los más significativos de estos once bists se pueden descartar dejando una potencia de salida de corrección sinx/x que tenga diez bits. Inclusive si este bit más significativo no se descartare, la potencia de salida del filtro de corrección de sinx/x (20) no excederá de diez bits cuando el filtro de corrección sinx/x (20) reciba los datos del símbolo normal. Por consiguiente, el bit más significativo se puede descartar sin riesgo de sobreflujo. La Tabla 1 siguiente muestra un conjunto ejemplar de valores de derivación para el filtro de corrección sinx/x (20).

Tabla 1

Derivación	Valor del Canal (3)	Valor del Canal (4)
0	- 3	- 4
1	1	2
2	- 22	- 30
3	400	400
4	- 22	- 30
5	1	2
6	-3	- 4

La respuesta de frecuencia del filtro de corrección sinx/x (20) basada en el conjunto ejemplar de los valores de

derivación del canal 3 se muestra en la Figura 5a, y la respuesta de frecuencia del filtro de corrección de sinx/x (20) basada en el conjunto ejemplar de los valores de corrección del canal 4 se muestra en la Figura 5b.

La potencia de salida del filtro de corrección sinx/x (20) es desplazada por 2.69 MHZ, que constituye una cuarta parte de la velocidad del símbolo f_{sym}. Este desplazamiento en frecuencia mueve el piloto DC al centro del espectro que se muestra en la Figura 4. El desplazamiento de frecuencia se lleva a cabo multiplicando los datos del símbolo por $e^{j1/4\omega sym}$. Esta multiplicación se implementa en la Figura 2 por un multiplicador (22) que multiplica los datos por $\cos(\omega_{sym}n/4)$ y un multiplicador (24) que multiplica los datos por $\sin(\omega_{sym}n/4)$ porque

$$e^{\frac{jn\omega sym}{4}} = \cos(\underline{\omega_{sym}n}) + j\sin(\underline{\omega_{sym}n}) . \qquad (1)$$
$$4 \qquad\qquad 4$$

como el desplazamiento de frecuencia es igual a una cuarta parte de la velocidad del reloj del símbolo, el factor $\cos(\omega_{sym}n/4)$ aplicado por el multiplicador (22) produce la siguiente secuencia:

$$1,0,-1,0,1,0,-1,0 \ . \ . \ .$$

y el factor $\sin(\omega_{sym}n/4)$ aplicado por el multiplicador (24) produce la siguiente secuencia:

$$0,1,0,-1,0,1,0,-1 \ldots .$$

Las dos secuencias se pueden intercalar porque un valor si y un valor no es cero en ambas secuencias. La secuencia resultante es dada por lo siguiente:

$$1,1,-1,-1,1,1,-1,-1 \ldots .$$

Tal como se muestra en la Figura 3, esta secuencia combinada puede implementarse como un circuito de complemento de la 2 (26). La señal de control del bloque de temporización (28) activa el circuito de complemento de la 2 (26) por ciclos de reloj de dos símbolos y la desactiva por ciclos de reloj de símbolos.

Una vez que la frecuencia de datos se ha desplazado, las partes reales e imaginarias de los datos se filtran mediante filtros de paso bajo (30 y 32), tal como se muestra en la Figura 2. Como aparece en la Figura 3, los filtros de bajo paso (30 y 32) se pueden implementar, por ejemplo, como un sólo filtro de paso bajo de treinta y tres desviaciones (34) porque los datos reales e imaginarios están intercalados en la potencia de salida del circuito de complemento de la 2

(26). El filtro de bajo paso (34) configura el espectro de potencia de salida VSB y tiene una atenuación progresiva del filtro que se aproxima a la atenuación progresiva del coseno elevado a la raíz del espectro de potencia de salida VSB ideal. El punto 3 db del filtro de paso bajo (34) está en 2.69 MHZ. Con el fin de reducir el tamaño del filtro de paso bajo (34), el ancho de banda del filtro de paso bajo (34) es ligeramente más ancho que el ancho de banda 3 MHZ de un filtro de coseno elevado a la raíz ideal. La Tabla 2 muestra un conjunto ejemplar de valores de desviación para el filtro de paso bajo (34), asumiendo que se implemente como un filtro de treinta y tres derivaciones.

Tabla 2

Derivación	Valor	Derivación	Valor	Derivación	Valor
0	- 5	11	66	22	- 20
1	- 14	12	22	23	- 41
2	- 7	13	122	24	17
3	14	14	- 24	25	27
4	11	15	384	26	- 14
5	- 18	16	631	27	- 18
6	- 14	17	384	28	11
7	27	18	- 24	29	14
8	17	19	- 122	30	- 7
9	- 41	20	22	31	- 14
10	- 20	21	66	32	- 5

La Figura 6 muestra la respuesta de frecuencia del filtro de paso bajo (34) que se basa en este conjunto ejemplar de treinta y tres valores de derivación.

Tal como se discutió previamente, los datos reales e imaginarios están intercalados en la potencia de salida del circuito de complemento de la 2 (26). Cuando los datos reales están en derivaciones pares del filtro de paso bajo (34), los datos imaginarios están en las derivaciones impares del filtro de paso bajo (34), y viceversa. Por consiguiente, el filtro de paso bajo (34) tiene dos potencias de salida, una potencia de salida de derivaciones pares y una potencia de salida de derivaciones impares. Ambas potencias de salida están intercaladas con los datos reales e imaginarios, y estas salidas se pueden redondear a once bits y se limitan a diez bits antes de que se apliquen a un multiplexor de dos a uno (36). Dicha limitación ocurre raramente.

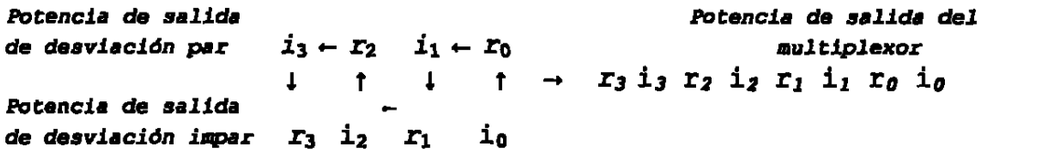
Con el fin de modular los datos, la velocidad de muestra se debe incrementar a 86.08 MHZ, que es ocho veces la velocidad de los datos del símbolo. La velocidad de la muestra se aumenta en tres etapas, con cada etapa interpolando los datos por dos. Cada etapa de interpolación incluye un filtro de media banda que remueve la imagen de los datos causada por la interpolación. Se requieren menos derivaciones de filtro si la interpolación y el filtrado se realizan en tres etapas en vez de una.

Por consiguiente, tal como se muestra en la Figura 2, las potencias de salida de los filtros de bajo paso (30 y 32) se suministran a la interpolación correspondiente y a los filtros de media banda (38 y 40) que realizan una interpolación por dos y que filtran los resultados de la interpolación. Las potencias de salida de la interpolación y de los filtros de media banda (38 y 40) se suministran luego a la interpolación correspondiente y a los filtros de media banda (42 y 44) que también realizan una interpolación por dos y que filtran los resultados de interpolación, y las potencias de salida de la interpolación y de los filtros de media banda (42 y 44) se suministran a la interpolación correspondiente y a los filtros de media banda (46 y 48) que también realizan una interpolación por dos y filtran los resultados de interpolación.

La interpolación por dos inserta un cero entre cada símbolo de datos complejo. Las partes reales e imaginarias de los símbolos de datos se intercalan luego porque un símbolo si y un símbolo no es cero. Por consiguiente, tal como se muestra en la Figura 3, solamente un filtro de banda media (50) y el multiplexor de dos a uno (36) se requieren para la interpolación de los filtros de media banda (38 y 40), sólo un filtro de media banda (52) y un multiplexor de dos a uno

(56) se requieren para la interpolación y para los filtros de media banda (42 y 44) y solamente un filtro de media banda (54) y un multiplexor de dos a uno (58) se requieren para la interpolación y para los filtros de media banda (46 y 48).

El multiplexor dos a uno (36), bajo control de bloque de temporización (28), intercala los datos combinando la potencias de salida de las derivaciones pares e impares del filtro de paso bajo (34) en un flujo de datos con los datos imaginarios intercalados antes de los datos reales. El siguiente diagrama ilustra los datos antes y después del multiplexor de dos a uno (36).



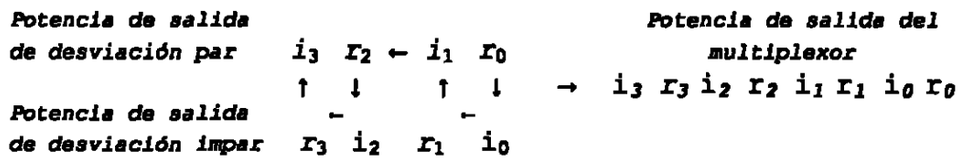
El filtro de banda media (50) se implementa, por ejemplo, como un filtro que tiene once desviaciones. La Tabla 3 siguiente muestra un conjunto ejemplar de valores de desviación para el filtro de media banda (50), asumiendo que se implementa como un filtro de once desviaciones.

Tabla 3

Desviación	Valor	Desviación	Valor
0	29	6	619
1	0	7	0
2	- 134	8	- 134
3	0	9	0
4	619	10	29
5	1024		

La Figura 7a muestra la respuesta de frecuencia para el filtro de media banda (50) que se basa en este conjunto de valores de desviación ejemplares.

El multiplexor dos a uno (56), bajo control del bloque de temporización (28), intercala los datos del filtro de media banda (50) combinando las potencias de salida de desviación pares e impares que salen del filtro de media banda (50) en un flujo de datos con los datos imaginarios intercalados después de los datos reales. El diagrama siguiente ilustra los datos antes y después del multiplexor de dos a uno (56):



El filtro de media banda (52) se implementa, por ejemplo, como un filtro que tiene siete desviaciones. La Tabla 4 siguiente muestra un conjunto ejemplar de valores de

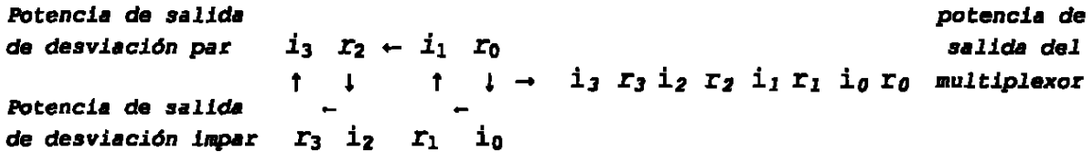
desviación para el filtro de media banda (52), asumiendo que se implementa como un filtro de siete desviaciones.

Tabla 4

Desviación	Valor	Desviación	Valor
0	- 9	4	73
1	0	5	0
2	73	6	- 9
3	128		

La Figura 7b muestra la respuesta de frecuencia para el filtro de media banda (52) que se basa en este conjunto de valores de desviación ejemplar.

El multiplexor de dos a uno (58), bajo control del bloque de temporización (28), intercala los datos del filtro de banda media (52) combinando las potencias de salida de desviación pares e impares del filtro de media banda (52) hacia un flujo de datos con los datos imaginarios intercalados después de los datos reales. El siguiente diagrama ilustra los datos antes y después del multiplexor de dos a uno (58).



El filtro de media banda (54) se implementa, por ejemplo, como un filtro que tiene siete desviaciones. La Tabla 5 siguiente muestra un conjunto ejemplar de valores de desviación para el filtro de media banda (54), asumiendo que se implementa como un filtro de siete desviaciones.

Tabla 5

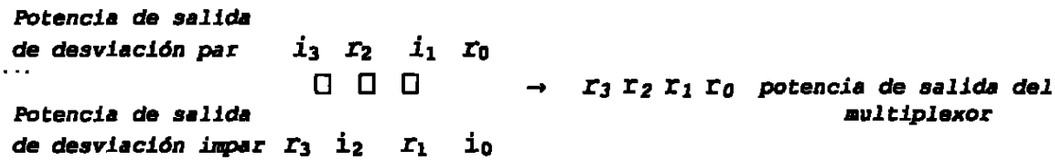
Desviación	Valor	desviación	Valor
0	- 1	4	9
1	0	5	0
2	9	6	- 1
3	16		

La Figura 7c muestra la respuesta de frecuencia para el filtro de media banda (54) que se basa en este conjunto de valores de desviación ejemplar.

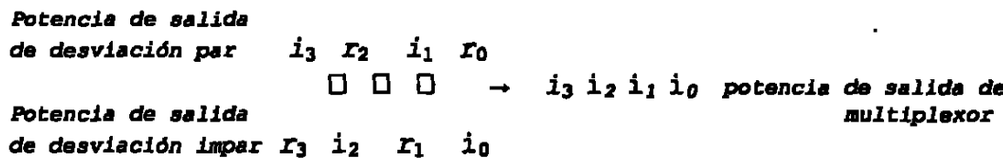
La respuesta de frecuencia de cada uno de los filtros de media banda (50, 52 y 54) es de 6 db por debajo a la mitad de la frecuencia Nyquist para cada uno de los filtros de media banda (50, 52 y 54). También, cada uno de los filtros de media banda (50, 52 y 54) tiene solamente una desviación impar. Como los datos reales e imaginarios están intercalados, los datos reales están en las desviaciones pares cuando los datos imaginarios están en las desviaciones impares y viceversa.

Por consiguiente, los filtros de media banda (50, 52 y 54) también tienen dos potencias de salida, una potencia de salida de desviación par y una potencia de salida de desviación impar. Ambas potencias de salida de cada uno de los filtros de media banda (50, 52 y 54) están intercaladas con los datos reales e imaginarios. Las potencias de salida de cada uno de los multiplexores dos a uno están redondeadas hasta once bits y limitadas a diez bits.

Los multiplexores dos-dos-a-uno (60 y 62) dispuestos en paralelo siguen al filtro de banda media (54). Los multiplexores dos a uno (60 y 62), bajo control del bloque de temporización (28), convierten los dos flujos de datos intercalados con datos reales e imaginarios en dos flujos de datos que son bien sea reales o imaginarios. El diagrama siguiente ilustra los datos antes y después del multiplexor de dos a uno (60).



Similarmente, el diagrama siguiente ilustra los datos antes y después del multiplexor dos a uno (62).



En el multiplexor dos a uno (36), los datos imaginarios se insertan antes de los datos reales. Esta operación hace que los datos reales e imaginarios sean sesgados por un ciclo de reloj. El multiplexor dos a uno (56) y el multiplexor dos a uno (58) insertan los datos reales antes de los datos imaginarios de manera que el sesgo del ciclo de un reloj no crezca. Un registro de diez bits (64) que se muestra en la Figura 3 se coloca después del multiplexor de dos a uno (62) con el fin de suministrar una demora de ciclo de reloj para remover este sesgo.

Un oscilador controlado numéricamente (66) tiene dos potencias de salida, una potencia de salida de onda de seno de ocho bits y una potencia de salida de onda de coseno de ocho bits. El acumulador de fases del oscilador controlado numéricamente (66) es de diez y seis bits. Para el canal 3, el incremento de fase es de 17.582, y la frecuencia de la potencia de salida es de 23.098348 MHZ. Para el canal 4, el incremento de fase es de 13.015, y la frecuencia de salida es

de 17.098453 MHZ. La frecuencia del piloto (f_{out}) se determina de acuerdo con la siguiente ecuación:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4} f_1)$$

en donde f_1 representa la frecuencia del símbolo, donde el factor $8f_1$ proviene de las tres interpolaciones por dos, en donde el factor $f_1/4$ proviene del circuito del complemento del 2 (26) y en donde el factor f_2 representa la potencia de salida del oscilador controlado numéricamente (66). Por consiguiente, la frecuencia del piloto para el canal 3 es de 60.308995 MHZ, y la frecuencia del piloto para el canal 4 es de 66.308890 MHZ. Como el acumulador de fases es solamente de diez y seis bits, las frecuencias del piloto están ligeramente por debajo de las frecuencias ideales de piloto. La señal de selección de canal controla la frecuencia NCO. Por ejemplo, un nivel lógico de cero selecciona el canal 3, y un nivel lógico de uno selecciona el canal 4.

Los datos complejos se modulan por la potencia de salida de la cuadratura del oscilador controlado numéricamente (66). Un multiplicador (68) multiplica el componente real de diez bits de los datos del símbolo que vienen del multiplexor dos a uno (60) por la potencia de salida del coseno de ocho bits del oscilador controlado numéricamente (66). Un multiplicador

(70) multiplica el componente imaginario de diez bits de los datos del símbolo proveniente del multiplexor dos a uno (62) y del registro de diez bits (64) por la potencia de salida con seno de ocho bits del oscilador controlado numéricamente (66). Un sumador (72) resta la potencia de salida del multiplicador (70) de la potencia de salida del multiplicador (68) que es equivalente a desplazar el espectro de la frecuencia de los datos complejos y su conjugado por la frecuencia NCO sumando las dos cifras. Esta operación se muestra en la siguiente ecuación:

$$X_n = \frac{(r_n + ji_n)e^{jn\omega_{nco}} + (r_n - ji_n)e^{-jn\omega_{nco}}}{2} \qquad (3)$$

$$= r_n \cos(\omega_{nco} n) - i_n \sin(\omega_{nco} n)$$

Los datos modulados tienen una frecuencia de centro igual a la frecuencia de la potencia de salida del oscilador controlado numéricamente (66). La frecuencia del piloto es de 2.69 MHZ por encima de la frecuencia de salida del oscilador controlado numéricamente (66). La Figura 8a muestra el espectro de frecuencia de los datos modulados en el canal 3, la Figura 8b muestra el especto de frecuencia de los datos modulados en el canal 4.

La potencia de salida del sumador (72) se atenúa por un atenuador (74) y se convierte a una señal análoga por un convertidor digital a análogo (76). La disminución debida a la atenuación progresiva del sinx/x del convertidor digital a análogo (76) puede ser superior para el canal 4 de lo que lo es para el canal 3. También, la ganancia DC del filtro de corrección sinx/x (20) puede ser ligeramente superior para el canal 3 de lo que lo es para el canal 4. Por consiguiente, si existen estas condiciones, la potencia de la señal después del convertidor digital a análogo (76) es superior (por ejemplo en alrededor de 3 db) para el canal 3 que para el canal 4. Por consiguiente, el atenuador (74) se proporciona para aplicar una atenuación adecuada (tal como 23/32) para hacer que la potencia de salida para el canal 3 y para el canal 4 sean sustancialmente iguales. Así pues, cuando la señal de selección de canal indica el canal 3, la atenuación aplicada por el atenuador (74) se activa y, cuando la señal de selección de canal indica el canal 4, la atenuación aplicada por el atenuador (74) se desactiva. El atenuador (74) también se puede arreglar para redondear los datos a nueve bits y para limitar los datos a ocho bits independientemente del canal.

El convertidor digital a análogo (76) tiene una resolución de ocho bits y una potencia de salida diferencial. Esta potencia de salida diferencial se convierte a una sola potencia de salida a través de un transformador 1:1 (78) del circuito de salida (16), tal como se muestra en la Figura 10. El costado del transformador 1:1 (78) acoplado al convertidor digital a análogo (76) tiene una desviación de centro a tierra y la impedancia de salida del convertidor digital a análogo (76) se establece colocando un resistor de 75Ω a lo largo de la salida del convertidor digital a análogo (76).

La Figura 9a muestra el especto de frecuencia en la salida del convertidor digital a análogo (76) para el canal 3, y la Figura 9b muestra el especto de frecuencia en la salida del convertidor digital a análogo (76) para el canal 4. La frecuencia del oscilador controlado numéricamente (66) se puede elegir de manera que la primera imagen de los datos modulados por encima de la frecuencia Nyquist recaigan sustancialmente en cualquiera de los canales 3 o 4. La corrección sinx/x impuesta por el filtro de corrección sinx/x (20) también se puede diseñar para corregir esta imagen. La relación de señal a ruido de la primera imagen también es lo suficientemente buena para ser recibida por un receptor. Por

consiguiente, la primera imagen se puede utilizar puesto que la potencia de salida modulada y la potencia de salida fundamental del convertidor digital a análogo (76) se puede ignorar.

Por consiguiente, un filtro de paso de banda (80) del circuito de salida (16), tal como se muestra en la Figura 10, se dispone para remover los datos modulados por debajo de los canales 3 y 4 y todas las imágenes por encima de los canales 3 y 4. Un atenuador (82) baja la potencia de la señal a una potencia nominal de 0 dbmv.

Arriba se han discutido algunas modificaciones y alternativas del presente invento. Otras modificaciones y alternativas ocurrirán para aquellos que practican el arte de lo aquí contemplado. De esta manera, la descripción del presente invento se debe interpretar solamente como una ilustración y tiene el objeto de enseñar a quienes conocen la técnica la mejor modalidad de llevar a cabo el invento. Los detalles se pueden variar sustancialmente sin apartarse del espíritu del invento, y se reserva la utilización exclusiva de las modificaciones que están dentro del alcance de las reivindicaciones adjuntas.

CODIFICADOR VSB Y MODULADOR DE RF DE BAJO COSTO

PARA SUMINISTRAR UNA SEÑAL VSB SUSTANCIALMENTE DE 6 MHZ VSB

A UN RECEPTOR DIGITAL DE TELEVISIÓN

REIVINDICACIONES:

1. Un sistema para suministrar una señal digital comprimida a un receptor de televisión digital, que comprende:

una fuente de señal de banda de base comprimida;

un codificador que codifica la señal de la banda de base en un formato digital, en donde la señal de la banda de base tiene una frecuencia;

un primer desplazador de frecuencia que desplaza la frecuencia de la señal de la banda de base codificada;

un primer interpolador que interpola una potencia de salida del primer desplazador de frecuencia;

un segundo interpolador que interpola una potencia de salida del primer interpolador;

un tercer interpolador que interpola una potencia de salida del segundo interpolador; y

un segundo desplazador de frecuencia que desplaza una frecuencia de una salida del tercer interpolador.

2. El sistema de la reivindicación 1 en donde el codificador proporciona la señal de la banda de base a una frecuencia f_1, y en donde el primer desplazador de frecuencia desplaza la señal de la banda de base en $f_1/4$.

3. El sistema de la reivindicación 2 en donde el primer interpolador interpola por dos, en donde el segundo interpolador interpola por dos y en donde el tercer interpolador interpola por dos.

4. El sistema de la reivindicación 2 en donde el primero, segundo y tercer interpoladores colectivamente interpolan por ocho.

5. El sistema de la reivindicación 4 en donde el primero y segundo desplazadores de frecuencia del primero, segundo y tercer interpoladores operan eficazmente de conformidad con la siguiente ecuación:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4} f_1)$$

en donde $8f_1$ proviene del primero, segundo y tercero interpolares, en donde $f_1/4$ proviene del primer desplazador de frecuencia, y en donde f_2 está suministrada por el segundo desplazador de frecuencia.

6. El sistema de la reivindicación 1 que comprende además un filtro de paso bajo acoplado entre el codificador y el primer interpolador, en donde el codificador proporciona la señal de la banda de base a una frecuencia f_1, y en donde el filtro de paso bajo tiene una respuesta que se desenvuelve aproximadamente en $f_1/4$.

7. El sistema de la reivindicación 6 en donde el desplazador de la primera frecuencia desplaza la señal de la banda de base en $f_1/4$.

8. El sistema de la reivindicación 7 en donde el primero, segundo y tercer interpoladores colectivamente interpolan por ocho.

9. El sistema de la reivindicación 8 en donde el primero y segundo desplazadores de frecuencia en el primero, segundo y tercer interpoladores operan con eficacia de acuerdo con la siguiente ecuación

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4} f_1)$$

en donde $8f_1$ proviene del primer, segundo y tercer interpoladores, donde $f_1/4$ proviene del primer desplazador de frecuencia, en donde f_2 es suministrada por el segundo desplazador de frecuencia.

10. El sistema de la reivindicación 1 que comprende además una corrección sinx/x acoplada entre el codificador y el primer interpolador, en donde la corrección sinx/x proporciona compensación para la atenuación progresiva sinx/x creada por el procesamiento en el sistema.

11. El sistema de la reivindicación 10 que comprende además un filtro de paso bajo acoplado entre el codificador y el primer interpolador, en donde el codificador proporciona la señal de la banda de base a la frecuencia f_1, y en donde el filtro de paso bajo tiene una respuesta que se atenúa progresivamente aproximadamente en $f_1/4$.

12. El sistema de la reivindicación 10 en donde el codificador proporciona la señal de la banda de base a una frecuencia f_1, y en donde el primer desplazador de frecuencia desplaza la señal de la banda de base en $f^1/4$.

13. El sistema de la reivindicación 12 en donde el primero, segundo y tercer interpoladores colectivamente interpolan por ocho.

14. El sistema de la reivindicación 13 en donde el primero y segundo desplazadores de frecuencia y el primero,

segundo y tercer interpoladores operan eficazmente de conformidad con la siguiente ecuación:

$$f_{out} = 8 f_1 - (f_2 + \frac{1}{4} f_1)$$

en donde $8f_1$ proviene del primero, segundo y tercer interpoladores, donde $f_1/4$ proviene del primer desplazador de frecuencia, y en donde f_2 es suministrada por el segundo desplazador de frecuencia.

15. El sistema de la reivindicación 1 en donde el primer interpolador interpola por dos, el segundo interpolador interpola por dos y el tercer interpolador interpola por dos.

16. Un método de suministrar una señal digital comprimida a un receptor de televisión digital que comprende:

codificar una señal de banda de base en un formato digital, en donde la señal de base tiene una frecuencia de símbolos;

aumentar la frecuencia de símbolos de la señal de banda de base codificada para producir una señal de frecuencia intermedia, en donde el aumento de la frecuencia de símbolos de la señal de banda de base codificada incluye tres interpolaciones de la señal de banda de base codificada; y

modular la señal de frecuencia intermedia con una señal de frecuencia de selección de canal para producir una señal de potencia de salida que tenga la frecuencia de un canal seleccionado.

17. El método de la reivindicación 16 en donde la señal de la banda de base tiene una frecuencia de símbolo f_1, y en donde el aumento de la frecuencia de símbolo de la señal de la banda de base codificada comprende desplazar la frecuencia del símbolo de la señal de banda de base en $f_1/4$.

18. El método de la reivindicación 16 en donde las tres interpolaciones son cada una, una interpolación por dos.

19. El método de la reivindicación 16 en donde la señal de la banda de base tiene una frecuencia de símbolo f_1, y en donde la señal de la potencia de salida tiene una frecuencia de f_{out} dada por la siguiente ecuación:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4} f_1)$$

en donde $8f_1$ proviene de las tres interpolaciones, donde $f_1/4$ proviene de un desplazamiento de frecuencia de la señal de banda de base codificada, y donde f_2 es la frecuencia de la señal de frecuencia de selección de canal.

20. El método de la reivindicación 16 que comprende además filtrar con paso bajo la señal de banda de base codificada, en donde la señal de banda de base tiene una frecuencia de símbolo f_1 y en donde el filtrado de paso bajo tiene un respuesta que se desenvuelve aproximadamente en $f_1/4$.

21. El método de la reivindicación 20 en donde el aumento de la frecuencia de los símbolos de la señal de banda de base codificada comprende desplazar la frecuencia de símbolos de la señal de base en $f_1/4$.

22. El método de la reivindicación 20 en donde las tres interpolaciones son cada una, una interpolación por dos.

23. El método de la reivindicación 20 en donde la señal de potencia de salida tiene una frecuencia f_{out} dada por la siguiente ecuación:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4} f_1)$$

en donde $8f_1$ proviene de las tres interpolaciones, en donde $f_1/4$ proviene del desplazamiento de frecuencia de la señal de banda de base codificada, y en donde f_2 es la frecuencia de la señal de frecuencia de selección de canal.

24. El método de la reivindicación 16 que comprende además la corrección sinx/x de la señal de banda de base codificada, en donde la corrección sinx/x compensa la atenuación progresiva sinx/x creada por el procesamiento del flujo descendente de corrección sinx/x.

25. El método de la reivindicación 24 que comprende además filtrar con paso bajo la señal de banda de base codificada, en donde la señal de banda de base tiene una frecuencia de símbolo f_1, y en donde el filtrado de paso bajo tiene una respuesta que se atenúa progresivamente aproximadamente en $f_1/4$.

26. El método de la reivindicación 24 en donde la señal de la banda de base tiene una frecuencia de símbolo f_1, y en donde el aumento de la frecuencia del símbolo de la señal de la banda de base codificada comprende desplazar la frecuencia del símbolo de la señal de banda de base en $f_1/4$.

27. El método de la reivindicación 24 en donde las tres interpolaciones son cada una, una interpolación por dos.

28. El método de la reivindicación 24 en donde la señal de la banda de base tiene una frecuencia de símbolo f_1, y en

donde la señal de la potencia de salida tiene una frecuencia f_{out} dada por la siguiente ecuación:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4} f_1)$$

en donde $8f_1$ proviene de las tres interpolaciones, donde $f_1/4$ proviene de un desplazamiento de frecuencia de la señal de banda de base codificada, y en donde f_2 es la frecuencia de la señal de frecuencia de selección de canal.

29. El método de la reivindicación 16 en donde la señal de potencia de salida contiene datos modulados por debajo del canal seleccionado y una imagen en el canal seleccionado, y en donde el método comprende el filtrado de la señal de salida para remover los datos modulados y para pasar la imagen.

30. Una señal eléctrica producida por (i) codificar una señal en un formato de televisión digital, en donde la señal tiene una frecuencia de símbolo f_1, (ii) desplazar la frecuencia de la señal codificada en $f_1/4$, (iii) aumentar la frecuencia de la señal codificada desplazada para producir una señal intermedia, y (iv) modular la señal de frecuencia intermedia con una señal de frecuencia de selección de canal de manera que se produzca una señal eléctrica en la frecuencia de un canal seleccionado.

31. La señal eléctrica de la reivindicación 30 en donde la frecuencia de la señal codificada desplazada es aumentada por interpolación, y en donde la interpolación comprende tres interpolaciones por dos.

32. La señal eléctrica de la reivindicación 30 que tiene una frecuencia de f_{out} dada por la siguiente ecuación:

$$f_{out} = 8 f_1 - (f_2 + \frac{1}{4} f_1)$$

en donde $8f_1$ proviene del aumento de la frecuencia de la señal codificada desplazada, en donde $f_1/4$ proviene del desplazamiento de frecuencia, y donde f_2 es la frecuencia de la señal de frecuencia de selección de canal.

33. La señal eléctrica de la reivindicación 30 en donde la señal codificada es filtrada con paso bajo, y en donde el filtrado con paso bajo tiene una respuesta que se atenúa progresivamente aproximadamente en $f_1/4$.

34. La señal eléctrica de la reivindicación 33 en donde la frecuencia de la señal codificada desplazada se aumenta por la interpolación, y en donde la interpolación comprende tres interpolaciones por dos.

35. La señal eléctrica de la reivindicación 33 que tiene una frecuencia f_{out} dada por la siguiente ecuación:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4} f_1)$$

en donde $8f_1$ proviene del aumento de la frecuencia de la señal codificada desplazada, en donde $f_1/4$ proviene de la frecuencia desplazada, y donde f_2 es la frecuencia de la señal de frecuencia de selección de canal.

36. La señal eléctrica de la reivindicación 30 en donde la señal codificada se corrige sinx/x, y en donde la corrección sinx/x compensa el desvolviendo sinx/x creado por el procesamiento de la señal eléctrica del flujo descendente de la corrección sinx/x.

37. La señal eléctrica de la reivindicación 36 en donde la señal codificada es filtrada con paso bajo, y en donde el filtrado de paso bajo tiene una respuesta que se desenvuelve aproximadamente a $f_1/4$.

38. La señal eléctrica de la reivindicación 36 en donde la frecuencia de la señal codificada desplazada es aumentada por la interpolación, y en donde la interpolación comprende tres interpolaciones por dos.

39. La señal eléctrica de la reivindicación 36 que tiene una frecuencia f_{out} dada por la siguiente ecuación:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4} f_1)$$

en donde $8f_1$ proviene del aumento de la frecuencia de la señal codificada desplazada, en donde $f_1/4$ proviene del desplazamiento de frecuencia, y donde f_2 es la frecuencia de la señal de frecuencia de selección de canal.

40. La señal eléctrica de la reivindicación 30 producida además por filtrado de manera que se remuevan los datos por debajo del canal seleccionado y que se deje una imagen de los datos como señal eléctrica.

RESUMEN DEL INVENTO

Un sistema suministra una señal digital comprimida a un receptor de televisión digital e incluye una fuente de señal de banda de base comprimida. Un codificador codifica la señal de la banda de base en un formato digital, y la señal de la banda de base tiene una frecuencia. La frecuencia de la señal de la banda de base codificada es desplazada por un primer desplazador de frecuencia. Un primer interpolador interpola la potencia de salida del primer desplazador de frecuencia, un segundo interpolador que interpola una frecuencia de salida del primer interpolador, y un tercer interpolador que interpola la potencia de salida del segundo interpolador. Un

segundo desplazador de frecuencia desplaza una frecuencia de una potencia de salida del tercer interpolador para producir una señal en un canal seleccionado.

102

PCT

REQUEST

The undersigned requests that the present international application be processed according to the Patent Cooperation Treaty.

For receiving Office use only

International Application No.

International Filing Date

Name of receiving Office and "PCT International Application"

Applicant's or agent's file reference *(if desired) (12 characters maximum)* 29471/36818

Box No. I TITLE OF INVENTION

LOW COST VSB ENCODER AND RF MODULATOR FOR SUPPLYING A SUBSTANTIALLY 6 MHZ VSB SIGNAL TO DIGITAL TELEVISION RECEIVER

Box No. II APPLICANT ☐ This person is also inventor

Name and address: *(Family name followed by given name; for a legal entity, full official designation. The address must include postal code and name of country. The country of the address indicated in this Box is the applicant's State (that is, country) of residence if no State of residence is indicated below.)*

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This person is applicant for the purposes of: ☒ all designated States ☐ all designated States except the United States of America ☐ the United States of America only ☐ the States indicated in the Supplemental Box

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☐ applicant and inventor
☒ inventor only *(If this check-box is marked, do not fill in below.)*

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State *(that is, country)* of nationality:

State *(that is, country)* of residence:

This person is applicant for the purposes of: ☐ all designated States ☐ all designated States except the United States of America ☐ the United States of America only ☐ the States indicated in the Supplemental Box

☐ Further applicants and/or (further) inventors are indicated on a continuation sheet.

Box No. IV AGENT OR COMMON REPRESENTATIVE; OR ADDRESS FOR CORRESPONDENCE

The person identified below is hereby/has been appointed to act on behalf of the applicant(s) before the competent International Authorities as: ☒ agent ☐ common representative

Name and address: *(Family name followed by given name; for a legal entity, full official designation. The address must include postal code and name of country.)*

JOIKE, Trevor B.
MARSHALL, O'TOOLE, GERSTEIN, MURRAY & BORUN
6300 Sears Tower
233 S. Wacker Drive
Chicago, Illinois 60606
United States of America

Telephone No.
312/474-6300

Facsimile No.
312/474-0448

Teleprinter No.

Agent's registration No. with the Office
25,542

☐ Address for correspondence: Mark this check-box where no agent or common representative is/has been appointed and the space above is used instead to indicate a special address to which correspondence should be sent.

Box No.V DESIGNATION OF STATES *Mark the applicable check-boxes below: at least one must be marked.*

The following designations are hereby made under Rule 4.9(a):

Regional Patent

☒ **AP ARIPO Patent:** GH Ghana, GM Gambia, KE Kenya, LS Lesotho, MW Malawi, MZ Mozambique, SD Sudan, SL Sierra Leone, SZ Swaziland, TZ United Republic of Tanzania, UG Uganda, ZW Zimbabwe, and any other State which is a Contracting State of the Harare Protocol and of the PCT

☒ **EA Eurasian Patent:** AM Armenia, AZ Azerbaijan, BY Belarus, KG Kyrgyzstan, KZ Kazakhstan, MD Republic of Moldova, RU Russian Federation, TJ Tajikistan, TM Turkmenistan, and any other State which is a Contracting State of the Eurasian

☒ **EP European Patent:** AT Austria, BE Belgium, CH & LI Switzerland and Liechtenstein, CY Cyprus, DE Germany, DK Denmark, ES Spain, FI Finland, FR France, GB United Kingdom, GR Greece, IE Ireland, IT Italy, LU Luxembourg, MC Monaco, NL Netherlands, PT Portugal, SE Sweden, TR Turkey, and any other State which is a Contracting State of the European Patent Convention and of the PCT

☒ **OA OAPI Patent:** BF Burkina Faso, BJ Benin, CF Central African Republic, CG Congo, CI Côte d'Ivoire, CM Cameroon, GA Gabon, GN Guinea, GW Guinea-Bissau, ML Mali, MR Mauritania, NE Niger, SN Senegal, TD Chad, TG Togo, and any other State which is a member State of OAPI and a Contracting State of the PCT *(if other kind of protection or treatment desired, specify on dotted line)* .....

National Patent *(if other kind of protection or treatment desired, specify on dotted line):*

☒	AE United Arab Emirates	☒	GE Georgia	☒	MW Malawi		
☒	AG Antigua and Barbuda	☒	GH Ghana	☒	MX Mexico		
☒	AL Albania	☒	GM Gambia	☒	MZ Mozambique		
☒	AM Armenia	☒	HR Croatia	☒	NO Norway		
☒	AT Austria	☒	HU Hungary	☒	NZ New Zealand		
☒	AU Australia	☒	ID Indonesia	☒	PL Poland		
☒	AZ Azerbaijan	☒	IL Israel	☒	PT Portugal		
☒	BA Bosnia and Herzegovina	☒	IN India	☒	RO Romania		
	.....	☒	IS Iceland	☒	RU Russian Federation		
☒	BB Barbados	☒	JP Japan		.....		
☒	BG Bulgaria	☒	KE Kenya	☒	SD Sudan		
☒	BR Brazil	☒	KG Kyrgyzstan	☒	SE Sweden		
☒	BY Belarus	☒	KP Democratic People's Republic	☒	SG Singapore		
☒	BZ Belize		of Korea	☒	SI Slovenia		
☒	CA Canada	☒	KR Republic of Korea	☒	SK Slovakia		
☒	CH & LI Switzerland and Liechtenstein	☒	KZ Kazakhstan	☒	SL Sierra Leone		
☒	CN China	☒	LC Saint Lucia	☒	TJ Tajikistan		
☒	CO Colombia	☒	LK Sri Lanka	☒	TM Turkmenistan		
☒	CR Costa Rica	☒	LR Liberia	☒	TR Turkey		
☒	CU Cuba	☒	LS Lesotho	☒	TT Trinidad and Tobago		
☒	CZ Czech Republic	☒	LT Lithuania		.....		
☒	DE Germany	☒	LU Luxembourg	☒	TZ United Republic of Tanzania		
☒	DK Denmark	☒	LV Latvia	☒	UA Ukraine		
☒	DM Dominica	☒	MA Morocco	☒	UG Uganda		
☒	DZ Algeria	☒	MD Republic of Moldova	☐	US United States of America		
☒	EE Estonia		.....		.....		
☒	ES Spain	☒	MG Madagascar	☒	UZ Uzbekistan		
☒	FI Finland	☒	MK The former Yugoslav Republic of	☒	VN Viet Nam		
☒	GB United Kingdom		Macedonia	☒	YU Yugoslavia		
☒	GD Grenada	☒	MN Mongolia	☒	ZA South Africa		
				☒	ZW Zimbabwe		

Check-boxes below reserved for designating States which have become party to the PCT after issuance of this sheet:

☒	EC Ecuador	☐	.....	☐	.....		
☒	GQ Equatorial Guinea	☐	.....	☐	.....		

Precautionary Designation Statement: In addition to the designations made above, the applicant also makes under Rule 4.9(b) all other designations which would be permitted under the PCT except any designation(s) indicated in the Supplemental Box as being excluded from the scope of this statement. The applicant declares that those additional designations are subject to confirmation and that any designation which is not confirmed before the expiration of 15 months from the priority date is to be regarded as withdrawn by the applicant at the expiration of that time limit. *(Confirmation (including fees) must reach the receiving Office within the 15-month time limit.)*

104

Box No. VI PRIORITY CLAIM

The priority of the following earlier application(s) is hereby claimed:

Filing date of earlier application *(day/month/year)*	Number of earlier application	Where earlier application is:		
		national application: country	regional application:* regional Office	international application: receiving Office
item (1) 04 August 2000 (04.08.00)	09/632,545	US		
item (2)				
item (3)				
item (4)				
item (5)				

☐ Further priority claims are indicated in the Supplemental Box.

The receiving Office is requested to prepare and transmit to the International Bureau a certified copy of the earlier application(s) *(only if the earlier application was filed with the Office which for the purposes of this international application is the receiving Office)* identified above as:

☐ all items ☒ item (1) ☐ item (2) ☐ item (3) ☐ item (4) ☐ item (5) ☐ other, see Supplemental Box

** Where the earlier application is an ARIPO application, indicate at least one country party to the Paris Convention for the Protection of Industrial Property or one Member of the World Trade Organization for which that earlier application was filed (Rule 4.10(b)(ii)):* .....

.....

Box No. VII INTERNATIONAL SEARCHING AUTHORITY

Choice of International Searching Authority (ISA) *(if two or more International Searching Authorities are competent to carry out the international search, indicate the Authority chosen; the two-letter code may be used):*

ISA/ EP.....

Request to use results of earlier search; reference to that search *(if an earlier search has been carried out by or requested from the International Searching Authority):*

Date *(day/month/year)* Number Country *(or regional Office)*

Box No. VIII DECLARATIONS

The following **declarations** are contained in Boxes Nos. VIII (i) to (v) *(mark the applicable check-boxes below and indicate in the right column the number of each type of declaration)*: Number of declarations

☐ Box No. VIII (i)	Declaration as to the identity of the inventor	:	
☐ Box No. VIII (ii)	Declaration as to the applicant's entitlement, as at the international filing date, to apply for and be granted a patent	:	
☐ Box No. VIII (iii)	Declaration as to the applicant's entitlement, as at the international filing date, to claim the priority of the earlier application	:	
☐ Box No. VIII (iv)	Declaration of inventorship (only for the purposes of the designation of the United States of America)	:	
☐ Box No. VIII (v)	Declaration as to non-prejudicial disclosures or exceptions to lack of novelty	:	

Box No. IX CHECK LIST; LANGUAGE OF FILING

This international application contains:

(a) the following number of sheets in paper form:

request (including declaration sheets)	:	4
description (excluding sequence listing part)	:	26
claims	:	14
abstract	:	1
drawings	:	7
Sub-total number of sheets	:	52
sequence listing part of description *(actual number of sheets if filed in paper form, whether or not also filed in computer readable form; see (b) below)*	:	
Total number of sheets	:	52

(b) sequence listing part of description filed in computer readable form

(i) ☐ only (under Section 801 (a)(i))

(ii) ☐ in addition to being filed in paper form (under Section 801 (a)(ii))

Type and number of carriers (diskette, CD-ROM, CD-R or other) on which the sequence listing part is contained *(additional copies to be indicated under item 9(ii), in right column)*: .....

This international application is accompanied by the following item(s) *(mark the applicable check-boxes below and indicate in right column the number of each item)*:

		Number of items
1. ☒ fee calculation sheet	:	1
2. ☐ original separate power of attorney	:	
3. ☐ original general power of attorney	:	
4. ☒ copy of general power of attorney; reference number, if any:	:	1
5. ☐ statement explaining lack of signature	:	
6. ☐ priority document(s) identified in Box No. VI as item(s):	:	
7. ☐ translation of international application into *(language)*:	:	
8. ☐ separate indications concerning deposited microorganism or other biological material	:	
9. ☐ sequence listing in computer readable form (indicate also type and number of carriers (diskette, CD-ROM, CD-R or other))		
(i) ☐ copy submitted for the purposes of international search under Rule 13*ter* only (and not as part of the international application)	:	
(ii) ☐ *(only where check-box (b)(i) or (b)(ii) is marked in left column)* additional copies including, where applicable, the copy for the purposes of international search under Rule 13*ter*	:	
(iii) ☐ together with relevant statement as to the identity of the copy or copies with the sequence listing part mentioned in left column	:	
10. ☒ other *(specify)*: postcard	:	1

Figure of the drawings which should accompany the abstract:

Language of filing of the international application: English

Box No. X SIGNATURE OF APPLICANT, AGENT OR COMMON REPRESENTATIVE

Next to each signature, indicate the name of the person signing and the capacity in which the person signs (if such capacity is not obvious from reading the request).

By _____

For receiving Office use only

1. Date of actual receipt of the purported international application:	2. Drawings:
3. Corrected date of actual receipt due to later but timely received papers or drawings completing the purported international application:	☐ received:
4. Date of timely receipt of the required corrections under PCT Article 11(2):	☐ not received:
5. International Searching Authority (if two or more are competent): ISA/	6. ☐ Transmittal of search copy delayed until search fee is paid

For International Bureau use only

Date of receipt of the record copy by the International Bureau:

This sheet is not part of and does not count as a sheet of the international application.

106

PCT

FEE CALCULATION SHEET

Annex to the Request

For receiving Office use only	
International Application No.	107
Date stamp of the receiving Office	

Applicant's or agent's file reference: **29471/36818**

Applicant: **ZENITH ELECTRONICS CORPORATION**

CALCULATION OF PRESCRIBED FEES

1. TRANSMITTAL FEE	240.00 T	
2. SEARCH FEE	846.00 S	

International search to be carried out by: EP

(If two or more International Searching Authorities are competent to carry out the international search, indicate the name of the Authority which is chosen to carry out the international search.)

3. INTERNATIONAL FEE

Basic Fee

Where item (b) of Box No. IX applies, enter Sub-total number of sheets
Where item (b) of Box No. IX does not apply, enter Total number of sheets: 82

b1	first 30 sheets		382.00 b1	
b2	22 (number of sheets)	x 9.00 (fee per sheet)	= 198.00 b2	
b3	additional component (only if sequence listing part of description is filed in computer readable form under Section 801(a)(i), or both in that form and on paper, under Section 801(a)(ii)): 400 x _____ (fee per sheet)		= b3	
	Add amounts entered at b1, b2 and b3 and enter total at B		580.00 B	

Designation Fees

The international application contains 87 designations.

	6 (number of designation fees payable *(maximum 6)*)	x 82.00 (amount of designation fee)	= 492.00 D	
	Add amounts entered at B and D and enter total at I		1,072.00 I	

(Applicants from certain States are entitled to a reduction of 75% of the international fee. Where the applicant is (or all applicants are) so entitled, the total to be entered at I is 25% of the sum of the amounts entered at B and D.)

4. FEE FOR PRIORITY DOCUMENT *(if applicable)*	P	
5. TOTAL FEES PAYABLE	2,158.00 TOTAL	

Add amounts entered at T, S, I and P, and enter total in the TOTAL box

☐ The designation fees are not paid at this time.

MODE OF PAYMENT

☐ authorization to charge ☐ postal money order ☐ cash ☐ coupons

☒ cheque ☐ bank draft ☐ revenue stamps ☐ other *(specify)*:

AUTHORIZATION TO CHARGE (OR CREDIT) DEPOSIT ACCOUNT

(This mode of payment may not be available at all receiving Offices)

Receiving Office: RO/ US

☐ Authorization to charge the total fees indicated above.

Deposit Account No.: 13-2855

☒ *(This check-box may be marked only if the conditions for deposit accounts of the receiving Office so permit)* Authorization to charge any deficiency or credit any overpayment in the total fees indicated above.

Date: 23/07/01

Name: Trevor B. Lake

☒ Authorization to charge the fee for priority document.

Signature: Trevor B. Lake

See Notes to the fee calculation sheet

107

PCT

GENERAL POWER OF ATTORNEY

(for several international applications filed under the Patent Cooperation Treaty)

(PCT Rule 90.5)

The undersigned person(s):
(Family name followed by given name; for a full legal entity, full official designation. The address must include postal code and name of country.)

ZENITH ELECTRONICS CORPORATION
2000 Millbrook Drive
Lincolnshire, Illinois 60069
United States of America

hereby appoint(s) the following person as: ☒ **agent** ☐ **common representative**

Name and address
(Family name followed by given name; for a legal entity, full official designation. The address must include postal code and name of country.)

JOIKE, Trevor B. and HEPPERMANN, Roger A.
MARSHALL, O'TOOLE, GERSTEIN, MURRAY & BORUN
6300 Sears Tower
233 S. Wacker Drive
Chicago, Illinois 60606
United States of America

to represent the undersigned before

☒ **all the competent International Authorities**

☐ **the International Searching Authority only**

☐ **the International Preliminary Examining Authority only**

in connection with any and all international applications filed by the undersigned with the following Office

United States Patent and Trademark Office **as receiving Office**

and to make or receive payments on behalf of the undersigned.

Signature(s) ***(where there are several persons, each of them must sign; next to each signature, indicate the name of the person signing and the capacity in which the person signs, if such capacity is not obvious from reading this power):***

By: Jack Kail
Name: Jack Kail
Title: Assistant Secretary

Date: June 15, 2001

PATENT COOPERATION TREATY

108 ~~108~~

From the RECEIVING OFFICE

To:

TREVOR B. JOIKE
HARSHALL. O'TOOLE. GERSTEIN. MURRAY &
BORUN: 6300 SEARS TOWER
233 S. WACKER DRIVE
CHICAGO IL 60606

PCT

NOTIFICATION OF THE INTERNATIONAL APPLICATION NUMBER AND OF THE INTERNATIONAL FILING DATE

(PCT Rule 20.5(c))

Date of mailing (*day/month/year*)	10 AUG 2001
Applicant's or agent's file reference: 29471/36818	IMPORTANT NOTIFICATION

International application No.	International filing date (*day/month/year*)	Priority date (*day/month/year*)
PCT/US01/23211	23 JUL 01	04 AUG 00

Applicant: ZENITH ELECTRONICS CORPORATION

Title of the invention: LOW COST VSB ENCODER AND RF MODULATOR FOR SUPPLYING A SUBSTANTIALLY 6 MHZ VSB SIGNAL TO

1. The applicant is hereby notified that the international application has been accorded the international application number and the international filing date indicated above.

2. The applicant is further notified that the record copy of the international application:

☑ was transmitted to the International Bureau on 10 AUG 01.

☐ has not yet been transmitted to the International Bureau for the reason indicated below and a copy of this notification has been sent to the International Bureau*:

☐ because the necessary national security clearance has not yet been obtained.

☐ because (*reason to be specified*):

* The International Bureau monitors the transmittal of the record copy by the receiving Office and will notify the applicant (with Form PCT/IB/301) of its receipt. Should the record copy not have been received by the expiration of 14 months from the priority date, the International Bureau will notify the applicant (Rule 22.1(c)).

3. FOREIGN TRANSMITTAL LICENSE INFORMATION Completed by: [illegible]

☑ Additional license for foreign transmittal not required. This subject matter is covered by a license already granted on the equivalent U.S. national application. Refer to that license for information concerning its scope.

☐ License for foreign transmittal not required. 37 CFR 5.11(e)(1) or 37 CFR 5.11(e)(2). However, a license may be required for additional subject matter. See 37 CFR 5.15(b).

☐ Foreign transmittal license granted. 35 U.S.C. 184; 37 CFR 5.11 on 10/4/01 (date):

☐ 37 CFR 5.15(a) ☐ 37 CFR 5.15(b)

Name and mailing address of the receiving Office Assistant Commissioner for Patents Box PCT Washington, D.C. 20231 Attn: RO US Facsimile No.	Authorized officer Elena Rivera PCT Operations (PCT Team I [illegible] 703) 305-3230 FAX: Telephone No.

Form PCT RO 105 (July 1992)

109 ~~103~~

The demand must be filed directly with the competent International Preliminary Examining Authority or, if two or more Authorities are competent, with the one chosen by the applicant. The full name or two-letter code of that Authority may be indicated by the applicant on the line below:

IPEA/ EP

PCT

CHAPTER II

DEMAND

under Article 31 of the Patent Cooperation Treaty:
The undersigned requests that the international application specified below be the subject of international preliminary examination according to the Patent Cooperation Treaty and hereby elects all eligible States (except where otherwise indicated).

For International Preliminary Examining Authority use only	
Identification of IPEA	Date of receipt of DEMAND

Box No. I IDENTIFICATION OF THE INTERNATIONAL APPLICATION		Applicant's or agent's file reference
		FP01,0516WO 29471/36818
International application No.	International filing date *(day/month/year)*	(Earliest) Priority date *(day/month/year)*
PCT/US01/23211	23 July 2001 (23-07-01)	04 August 2000 (04-08-00)

Title of invention
LOW COST VSB ENCODER AND RF MODULATOR FOR SUPPLYING A SUBSTANTIALLY 6 MHZ VSB SIGNAL TO DIGITAL TELEVISION RECEIVER

Box No. II APPLICANT(S)

Name and address: *(Family name followed by given name; for a legal entity, full official designation. The address must include postal code and name of country.)*	Telephone No.
Zenith Electronics Corporation 2000 Millbrook Drive Lincolnshire, Illinois 60069 United States of America	847/941-8018
	Facsimile No. 847/941-8003
	Teleprinter No.
	Applicant's registration No. with the Office

State *(that is, country)* of nationality:	State *(that is, country)* of residence:
US	US

Name and address: *(Family name followed by given name; for a legal entity, full official designation. The address must include postal code and name of country.)*

State *(that is, country)* of nationality:	State *(that is, country)* of residence:

Name and address: *(Family name followed by given name; for a legal entity, full official designation. The address must include postal code and name of country.)*

State *(that is, country)* of nationality:	State *(that is, country)* of residence:

☐ Further applicants are indicated on a continuation sheet.

110

Sheet No. ...

International application No.
PCT/US01/23211

Box No. III AGENT OR COMMON REPRESENTATIVE; OR ADDRESS FOR CORRESPONDENCE

The following person is ☒ agent ☐ common representative

and ☒ has been appointed earlier and represents the applicant(s) also for international preliminary examination.

☒ is hereby appointed and any earlier appointment of (an) agent(s)/common representative is hereby revoked.

☐ is hereby appointed, specifically for the procedure before the International Preliminary Examining Authority, in addition to the agent(s)/common representative appointed earlier.

Name and address: *(Family name followed by given name; for a legal entity, full official designation. The address must include postal code and name of country.)*

JOIKE, Trevor B.
Schiff Hardin & Waite
6600 Sears Tower
Chicago, Illinois 60606
United States of America

Telephone No.
312/258-5774

Facsimile No.
312/258-5700

Teleprinter No.

Agent's registration No. with the Office
25,542

☐ **Address for correspondence:** Mark this check-box where no agent or common representative is/has been appointed and the space above is used instead to indicate a special address to which correspondence should be sent.

Box No. IV BASIS FOR INTERNATIONAL PRELIMINARY EXAMINATION

Statement concerning amendments:*

1. The applicant wishes the international preliminary examination to start on the basis of:

☒ the international application as originally filed

the description
☒ as originally filed
☐ as amended under Article 34

the claims
☒ as originally filed
☐ as amended under Article 19 (together with any accompanying statement)
☐ as amended under Article 34

the drawings
☒ as originally filed
☐ as amended under Article 34

2. ☐ The applicant wishes any amendment to the claims under Article 19 to be considered as reversed.

3. ☐ The applicant wishes the start of the international preliminary examination to be postponed until the expiration of 20 months from the priority date unless the International Preliminary Examining Authority receives a copy of any amendments made under Article 19 or a notice from the applicant that he does not wish to make such amendments (Rule 69.1(d)). *(This check-box may be marked only where the time limit under Article 19 has not yet expired.)*

* Where no check-box is marked, international preliminary examination will start on the basis of the international application as originally filed or, where a copy of amendments to the claims under Article 19 and/or amendments of the international application under Article 34 are received by the International Preliminary Examining Authority before it has begun to draw up a written opinion or the international preliminary examination report, as so amended.

Language for the purposes of international preliminary examination: English

☒ which is the language in which the international application was filed.
☐ which is the language of a translation furnished for the purposes of international search.
☒ which is the language of publication of the international application.
☐ which is the language of the translation (to be) furnished for the purposes of international preliminary examination.

Box No. V ELECTION OF STATES

The applicant hereby elects all eligible States *(that is, all States which have been designated and which are bound by Chapter II of the PCT)*

excluding the following States which the applicant wishes not to elect:

Sheet No. ...

International application No.
PCT/US01/23211

Box No. VI CHECK LIST

The demand is accompanied by the following elements, in the language referred to in Box No. IV, for the purposes of international preliminary examination:

			For International Preliminary Examining Authority use only: received	not received
1. translation of international application	:	sheets	☐	☐
2. amendments under Article 34	:	sheets	☐	☐
3. copy (or, where required, translation) of amendments under Article 19	:	sheets	☐	☐
4. copy (or, where required, translation) of statement under Article 19	:	sheets	☐	☐
5. letter	:	sheets	☐	☐
6. other *(specify)*	:	sheets	☐	☐

The demand is also accompanied by the item(s) marked below:

1. ☒ fee calculation sheet:
2. ☐ original separate power of attorney
3. ☐ original general power of attorney
4. ☒ copy of general power of attorney; reference number, if any:
5. ☐ statement explaining lack of signature
6. ☐ sequence listing in computer readable form
7. ☐ other *(specify)*:

Box No. VII SIGNATURE OF APPLICANT, AGENT OR COMMON REPRESENTATIVE

Next to each signature, indicate the name of the person signing and the capacity in which the person signs (if such capacity is not obvious from reading the demand).

Trevor B. Jolke, Agent

Date: 08/02/2002

For International Preliminary Examining Authority use only

1. Date of actual receipt of DEMAND:
2. Adjusted date of receipt of demand due to CORRECTIONS under Rule 60.1(b):
3. ☐ The date of receipt of the demand is AFTER the expiration of 19 months from the priority date and item 4 or 5, below, does not apply. ☐ The applicant has been informed accordingly.
4. ☐ The date of receipt of the demand is WITHIN the period of 19 months from the priority date as extended by virtue of Rule 80.5.
5. ☐ Although the date of receipt of the demand is after the expiration of 19 months from the priority date, the delay in arrival is EXCUSED pursuant to Rule 82.

For International Bureau use only

Demand received from IPEA on:

CHAPTER II

PCT

FEE CALCULATION SHEET

Annex to the Demand

International application No. PCT/US01/23211	For International Preliminary Examining Authority use only
Applicant's or agent's file reference FP01,0516WO 29471/36818	Date stamp of the IPEA

Applicant

Zenith Electronics Corporation

CALCULATION OF PRESCRIBED FEES

1. Preliminary examination fee	1553.00	P	
2. Handling fee *(Applicants from certain States are entitled to a reduction of 75% of the handling fee. Where the applicant is (or all applicants are) so entitled, the amount to be entered at H is 25% of the handling fee.)*	147.00	H	
3. Total of prescribed fees Add the amounts entered at P and H and enter total in the TOTAL box	1680.00 TOTAL		

MODE OF PAYMENT

- ☐ authorization to charge deposit account with the IPEA (see below)
- ☐ cheque
- ☐ postal money order
- ☒ bank draft
- ☐ cash
- ☐ revenue stamps
- ☐ coupons
- ☐ other *(specify)*:

AUTHORIZATION TO CHARGE (OR CREDIT) DEPOSIT ACCOUNT
(This mode of payment may not be available at all IPEAs)

IPEA/ _____

☐ Authorization to charge the total fees indicated above.

☐ *(This check-box may be marked only if the conditions for deposit accounts of the IPEA so permit)* Authorization to charge any deficiency or credit any overpayment in the total fees indicated above.

Deposit Account No.: _____

Date: _____

Name: _____

Signature: _____

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
14 February 2002 (14.02.2002)

PCT

(10) International Publication Number
WO 02/13534 A2

(51) International Patent Classification[7]: H04N 7/24

(21) International Application Number: PCT/US01/23211

(22) International Filing Date: 23 July 2001 (23.07.2001)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
09/632,545 4 August 2000 (04.08.2000) US

(71) Applicant: ZENITH ELECTRONICS CORPORATION [US/US]; 2000 Millbrook Drive, Lincolnshire, IL 60069 (US).

(72) Inventor: HORWITZ, Thomas, A.; 1236 Spring Creek Road, Elgin, IL 60120 (US).

(74) Agent: JOIKE, Trevor, B.; Marshall, O'Toole, Gerstein, Murray & Borun, 6300 Sears Tower, 233 S. Wacker Drive, Chicago, IL 60606 (US).

(81) Designated States *(national)*: AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NO, NZ, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TR, TT, TZ, UA, UG, UZ, VN, YU, ZA, ZW.

(84) Designated States *(regional)*: ARIPO patent (GH, GM, KE, LS, MW, MZ, SD, SL, SZ, TZ, UG, ZW), Eurasian patent (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European patent (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE, TR), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Published:
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(54) Title: LOW COST VSB ENCODER AND RF MODULATOR FOR SUPPLYING A SUBSTANTIALLY 6 MHZ VSB SIGNAL TO DIGITAL TELEVISION RECEIVER

(57) Abstract: A system supplies a compressed digital signal to a digital television receiver and includes a source of compressed baseband signal. An encoder encodes the baseband signal in a digital format, and the baseband signal has a frequency. The frequency of the encoded baseband signal is shifted by a first frequency shifter. A first interpolator interpolates an output of the first frequency shifter, a second interpolator that interpolates an output of the first interpolator, and a third interpolator that interpolates an output of the second interpolator. A second frequency shifter shifts a frequency of an output of the third interpolator to produce a signal at a selected channel.

LOW COST VSB ENCODER AND RF MODULATOR FOR SUPPLYING A SUBSTANTIALLY 6 MHZ VSB SIGNAL TO DIGITAL TELEVISION RECEIVER

Background of the Invention and Prior Art

This invention relates generally to digital VSB (vestigial side band) television receivers and specifically to a simple low cost system for coupling a digital VSB signal from any of a plurality of sources of a digital signal (such as an MPEG - Motion Picture Experts Group - digital signal) to a digital VSB television receiver. While the description is directed to a VSB digital format, it should be understood that the invention in its broader aspects is not limited to a particular digital format.

The recently adopted Digital Television Standard specifies a VSB subsystem having a terrestrial broadcast mode (8 VSB) and a high data rate mode (16 VSB). There are other VSB modes available, namely 2 VSB, 4 VSB and 8 VSB (non terrestrial). The 8 VSB terrestrial mode has the data carrying capability of 4 VSB. (The Trellis coding that is added for the terrestrial environment creates the additional modulation levels.) The digital data signal for video is MPEG encoded and for audio is Dolby AC-3 processed and must be subjected to decompression before application to conventional video and audio circuitry. For terrestrial

broadcasting, the data signal is: randomized; subjected to Reed-Solomon (R/S) type encoding for error correction; interleaved; Trellis encoded; multiplexed with segment sync and field sync; supplied with a DC pilot; subjected to pre-equalization filtering; modulated; and RF upconverted for transmission. The digital television receiver includes a tuner and a VSB demodulator for developing the baseband signal, which is in compressed form. The demodulated signal is applied to an equalizer for equalizing the received signal. The baseband signal is applied to a transport demultiplexer which directs the data to an appropriate MPEG decoder and a Dolby decoder for recovering the video and audio in proper form for application to video and audio processing circuitry. Digital signals may be supplied to digital receivers by over-the-air transmission sources. Digital signals from other sources, such as DVD (digital video disk) players, VCR's (video cassette recorders), PC's (personal computers), digital cable boxes, satellite receivers and the like will also be supplied to digital television receivers, as is the case for present analog television receivers.

In an analog system, it is well known that coupling the signal from a VCR or other video source to a baseband input of a television receiver (if one is available) yields results superior to those achieved by modulating the signal to RF channel 3 or RF channel 4 and supplying it to the tuner input of the television receiver. This baseband coupling, however, does not necessarily produce superior results in a digital environment in which case the distance between the baseband signal source and the television receiver must be kept relatively short. In an analog system, it is also well known to supply double sideband signals to a receiver.

In accordance with the invention described in U.S. Application Serial No. 09/300,299 filed April 27, 1999, a compressed baseband digital signal is encoded in a given format for digital transmission and is modulated onto an RF carrier for direct application via a cable or wireless link to the RF input of a digital television receiver. The RF signal, which is at a low power level and has a less-than-nominal bandpass, takes advantage of the front end signal processing that is built into the digital television receiver. Signal impairments, due to the less-than-nominal

bandpass characteristic and noise introduced during transmission, that are below a given threshold are corrected by the correction circuitry in the digital television receiver front end. The low signal power, the corrective circuitry for the digital signal, and the benign environment of the communications link enable components and circuits of much lower tolerance (and cost) to be used in the encoding and modulating process implemented by the other sources described above. The result is an modulator that is very low cost and very effective in coupling a digital signal to a digital television receiver over an RF channel.

The present invention is directed to another arrangement for providing a digital signal to a digital receiver.

Summary of the Invention

In accordance with a first aspect of the invention, a system for supplying a compressed digital signal to a digital television receiver comprises a source of compressed baseband signal, an encoder, first and second frequency shifters, and first, second, and third interpolators. The encoder encodes the baseband signal in a

digital format, wherein the baseband signal has a frequency. The first frequency shifter shifts the frequency of the encoded baseband signal. The first interpolator interpolates an output of the first frequency shifter, the second interpolator interpolates an output of the first interpolator, and the third interpolator interpolates an output of the second interpolator. The second frequency shifter shifts a frequency of an output of the third interpolator.

In accordance with another aspect of the invention, a method of supplying a compressed digital signal to a digital television receiver comprises the following: encoding a baseband signal in a digital format, wherein the baseband signal has a symbol frequency; increasing the symbol frequency of the encoded baseband signal to produce an intermediate frequency signal, wherein the increasing of the symbol frequency of the encoded baseband signal includes three interpolations of the encoded baseband signal; and, modulating the intermediate frequency signal with a channel selection frequency signal so as to produce an output signal having the frequency of a selected channel.

In accordance with still another aspect of the invention, an electrical signal is produced by (i) encoding a signal in a digital television format, wherein the signal has a symbol frequency f_1, (ii) shifting the frequency of the encoded signal by $f_1/4$, (iii) increasing the frequency of the shifted encoded signal to produce an intermediate signal, and (iv) modulating the intermediate frequency signal with a channel selection frequency signal so as to produce the electrical signal at the frequency of a selected channel.

Brief Description of the Drawings

These and other features and advantages of the invention will be apparent upon reading the following description in conjunction with the drawing, in which:

Figure 1 is a simplified block diagram of a digital remodulator in accordance with the present invention;

Figure 2 is a block diagram showing the VSB digital signal modulator of Figure 1 in additional detail;

Figure 3 is a more detailed block diagram of the VSB digital signal modulator of Figure 2;

Figure 4 is a curve illustrating the frequency spectrum of the encoded baseband signal supplied to the VSB digital signal modulator of Figures 2 and 3;

Figure 5a illustrates the frequency response of the sinx/x correction filter of Figures 2 and 3 for channel 3;

Figure 5b illustrates the frequency response of the sinx/x correction filter of Figures 2 and 3 for channel 4;

Figure 6 illustrates the frequency response of the low pass filters of Figures 2 and 3;

Figure 7a illustrates the frequency response of the half band filter HBF 1 of Figures 2 and 3;

Figure 7b illustrates the frequency response of the half band filter HBF 2 of Figures 2 and 3;

Figure 7c illustrates the frequency response of the half band filter HBF 3 of Figures 2 and 3;

Figure 8a illustrates the frequency spectrum of the modulated data for channel 3;

Figure 8b illustrates the frequency spectrum of the modulated data for channel 4;

Figure 9a illustrates the frequency spectrum of the D/A output for channel 3;

Figure 9b illustrates the frequency spectrum of the D/A output for channel 4; and,

Figure 10 is a block diagram of an output circuit for the VSB digital signal modulator of Figures 2 and 3.

Description of the Preferred Embodiment

As shown in Figure 1, a source 10 provides a digital MPEG encoded signal. The source 10 may be any of a number of different types of sources. For example, the source 10 may comprise a satellite receiver, a VCR, a DVD, a digital cable box, a PC, etc. The output of the source 10 is an MPEG (and Dolby AC-3) encoded digital signal at baseband frequency. This baseband signal may be applied directly to a baseband input of a digital television receiver. However, in accordance with the present invention, the baseband signal is applied to an ATSC encoder 12 which subjects the signal to interleaving, randomizing, R/S error encoding, and/or Trellis encoding. Field and segment syncs are added and, while not indicated, a pilot is inserted and equalization filtering is employed. (See ATSC

Standard A/53, ATSC Digital Television Standard for a detailed description of these operations.)

The ATSC encoder 12 supplies symbol data to a digital modulator 14 which remodulates the symbol data onto a selected RF channel, such as channel 3 or 4. Following remodulation, an output circuit 16 processes the remodulated data and supplies the processed remodulated data over an RF output to a digital receiver such as a digital television. The symbol data provided by the ATSC encoder 12 has a 10.76 MHZ data rate and its spectrum is flat across the entire frequency band. This data has a small DC offset which is used as a pilot, and the symbol data width is seven bits. Figure 4 shows the symbol data frequency spectrum taken with a 128 point discrete Fourier Transform (DFT).

As shown in Figures 2 and 3, a sinx/x correction filter 20 is provided at a front end of the digital modulator 14. The sinx/x correction filter 20 may have seven taps and is provided to compensate for the sinx/x roll off created by the D/A converter discussed below. Because channel 3 requires a different sinx/x correction than does channel 4, the sinx/x correction filter 20 contains two sets of tap values selected by the channel select signal. For

example, if the channel select signal is a logic level zero, the channel 3 taps are selected and, if the channel select signal is a logic level one, the channel 4 taps are selected.

The bits produced by application of these taps to the symbol data may be rounded to eleven bits and the most significant of these eleven bits may be discarded leaving a sinx/x correction output having ten bits. Even if this most significant bit were not discarded, the output of the sinx/x correction filter 20 will not exceed ten bits when the sinx/x correction filter 20 receives normal symbol data. Therefore, the most significant bit may be discarded without risk of overflow. Table 1 below shows an exemplary set of tap values for the sinx/x correction filter 20.

Table 1

Tap	Ch 3 Value	Ch 4 Value
0	-3	-4
1	1	2
2	-22	-30
3	400	400
4	-22	-30
5	1	2
6	-3	-4

The frequency response of the sinx/x correction filter 20 based upon the exemplary set of channel 3 tap values is shown in Figure 5a, and the frequency response of the sinx/x correction filter 20 based upon the exemplary set of channel 4 tap values is shown in Figure 5b.

The output of the sinx/x correction filter 20 is shifted by 2.69 MHZ, which is one-fourth of the symbol rate f_{sym}. This shift in frequency moves the DC pilot to the center of the spectrum shown in Figure 4. The frequency shift is performed by multiplying the symbol data by $e^{j1/4\omega sym}$. This multiplication is implemented in Figure 2 by a multiplier 22 which multiplies the data by $\cos(\omega_{sym}n/4)$

and a multiplier 24 which multiplies the data by $\sin(\omega_{sym}n/4)$ because

$$e^{\frac{j n \omega_{sym}}{4}} = \cos\left(\frac{\omega_{sym} n}{4}\right) + j\sin\left(\frac{\omega_{sym} n}{4}\right). \quad (1)$$

Because the frequency shift is equal to one-fourth the symbol clock rate, the factor $\cos(\omega_{sym}n/4)$ applied by the multiplier 22 produces the following sequence:

$$1,0,-1,0,1,0,-1,0 \ . \ . \ .$$

and the factor $\sin(\omega_{sym}n/4)$ applied by the multiplier 24 produces the following sequence:

$$0,1,0,-1,0,1,0,-1 \ . \ . \ . \ .$$

The two sequences can be interleaved because every other value in both sequences is zero. The resulting sequence is given by the following:

$$1,1,-1,-1,1,1,-1,-1 \ . \ . \ . \ .$$

As shown in Figure 3, this combined sequence may be implemented as a 2's complement circuit 26. The control signal from the timing block 28 turns on the 2's complement circuit 26 for two symbol clock cycles and off for two symbol clock cycles.

After the data is shifted in frequency, the real and imaginary parts of the data are filtered by low pass filters 30 and 32 as shown in Figure 2. As shown in Figure 3, the low pass filters 30 and 32 may be implemented, for example, as a single thirty-three tap low pass filter 34 because the real and imaginary data are interleaved at the output of the 2's complement circuit 26. The low pass filter 34 shapes the VSB output spectrum and has a filter roll-off that approximates the root-raised cosine roll-off of the ideal VSB output spectrum. The 3 db point of the low pass filter 34 is at 2.69 MHZ. In order to reduce the size of the low pass filter 34, the bandwidth of the low pass filter 34 is slightly wider than the 3 MHZ bandwidth of an ideal root-raised cosine filter. Table 2 below shows an exemplary set of tap values for the low pass filter 34, assuming that it is implemented as a thirty-three tap filter.

Table 2

Tap	Value	Tap	Value	Tap	Value
0	-5	11	66	22	-20
1	-14	12	22	23	-41
2	-7	13	-122	24	17
3	14	14	-24	25	27
4	11	15	384	26	-14
5	-18	16	631	27	-18
6	-14	17	384	28	11
7	27	18	-24	29	14
8	17	19	-122	30	-7
9	-41	20	22	31	-14
10	-20	21	66	32	-5

Figure 6 shows the frequency response of the low pass filter 34 based upon this exemplary set of thirty-three tap values.

As discussed previously, the real and imaginary data are interleaved at the output of the 2's complement circuit 26. When the real data are at the even taps of the low pass filter 34, the imaginary data are at the odd taps of the low pass filter 34, and vice versa. Therefore, the low pass filter 34 has two outputs, an even tap output and an odd tap output. Both outputs are interleaved with the real and imaginary data, and these outputs may be rounded to

eleven bits and limited to ten bits before being applied to a two-to-one mux 36. Such limiting will rarely occur.

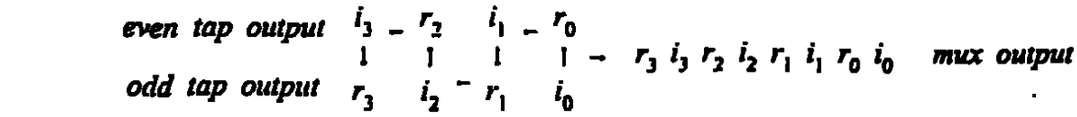
In order to modulate the data, the sample rate should be increased to 86.08 MHZ, which is eight times the symbol data rate. The sample rate is increased in three stages, with each stage interpolating the data by two. Each interpolation stage includes a half band filter which removes the image of the data caused by the interpolation. Fewer filter taps are required if the interpolation and filtering are done in three steps instead of one.

Accordingly, as shown in Figure 2, the outputs of the low pass filters 30 and 32 are supplied to corresponding interpolation and half band filters 38 and 40 which perform an interpolation by two and which filter the interpolation results. The outputs of the interpolation and half band filters 38 and 40 are then supplied to corresponding interpolation and half band filters 42 and 44 which also perform an interpolation by two and which filter the interpolation results, and the outputs of the interpolation and half band filters 42 and 44 are supplied to corresponding interpolation and half band filters 46 and 48

which likewise perform an interpolation by two and filter the interpolation results.

Interpolation by two inserts a zero in between each complex data symbol. The real and imaginary parts of the data symbols are then interleaved because every other data symbol is zero. Therefore, as shown in Figure 3, only one half band filter 50 and the two-to-one mux 36 are required for the interpolation and half band filters 38 and 40, only one half band filter 52 and a two-to-one mux 56 are required for the interpolation and half band filters 42 and 44, and only one half band filter 54 and a two-to-one mux 58 are required for the interpolation and half band filters 46 and 48.

The two-to-one mux 36, under control of the timing block 28, interleaves the data by combining the even and odd tap outputs of the low pass filter 34 into one data stream with the imaginary data interleaved before the real data. The diagram below illustrates the data before and after the two-to-one mux 36.



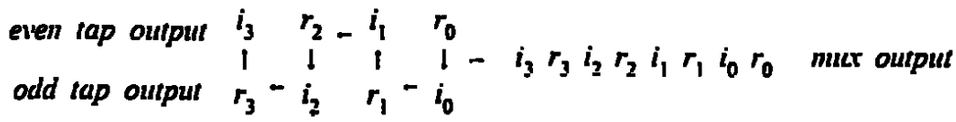
The half band filter 50 is implemented, for example, as a filter having eleven taps. Table 3 below shows an exemplary set of tap values for the half band filter 50, assuming that it is implemented as an eleven tap filter.

Table 3

Tap	Value	Tap	Value
0	29	6	619
1	0	7	0
2	-134	8	-134
3	0	9	0
4	619	10	29
5	1024		

Figure 7a shows the frequency response for the half band filter 50 based upon this set of exemplary tap values.

The two-to-one mux 56, under control of the timing block 28, interleaves the data from the half band filter 50 by combining the even and odd tap outputs of the half band filter 50 into one data stream with the imaginary data interleaved after real data. The diagram below illustrates the data before and after the two-to-one mux 56.



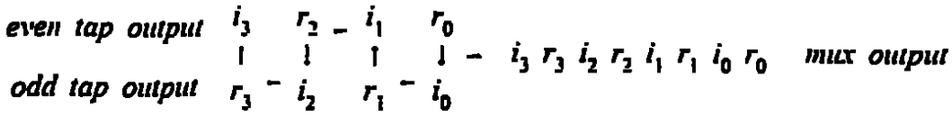
The half band filter 52 is implemented, for example, as a filter having seven taps. Table 4 below shows an exemplary set of tap values for the half band filter 52, assuming that it is implemented as an seven tap filter.

Table 4

Tap	Value	Tap	Value
0	-9	4	73
1	0	5	0
2	73	6	-9
3	128		

Figure 7b shows the frequency response for the half band filter 52 based upon this set of exemplary tap values.

The two-to-one mux 58, under control of the timing block 28, interleaves the data from the half band filter 52 by combining the even and odd tap outputs of the half band filter 52 into one data stream with the imaginary data interleaved after real data. The diagram below illustrates the data before and after the two-to-one mux 58.



The half band filter 54 is implemented, for example, as a filter having seven taps. Table 5 below shows an exemplary set of tap values for the half band filter 54, assuming that it is implemented as an seven tap filter.

Table 5

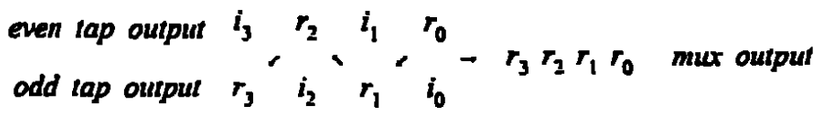
Tap	Value	Tap	Value
0	-1	4	9
1	0	5	0
2	9	6	-1
3	16		

Figure 7c shows the frequency response for the half band filter 54 based upon this set of exemplary tap values.

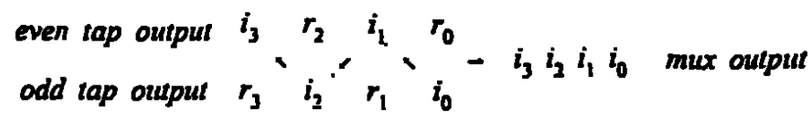
The frequency response of each of the half band filters 50, 52, and 54 is 6 db down at half of the Nyquist frequency for each of the half band filters 50, 52, and 54. Also, each of the half band filters 50, 52, and 54 has nly one odd tap. Because the real and imaginary data are interleaved, the real data are in the even taps when the

imaginary data are in the odd taps and vice versa. Therefore, the half band filters 50, 52, and 54 also have two outputs, an even tap output and an odd tap output. Both outputs of each of the half band filters 50, 52, and 54 are interleaved with the real and imaginary data. The outputs of each of the two-to-one multiplexers are rounded to eleven bits and limited to ten bits.

Two two-to-one muxes 60 and 62 arranged in parallel follow the half band filter 54. The two-to-one muxes 60 and 62, under control of the timing block 28, convert the two data streams interleaved with real and imaginary data to two data streams that are either real or imaginary. The diagram below illustrates the data before and after the two-to-one mux 60.



Similarly, the diagram below illustrates the data before and after the two-to-one mux 62.



At the two-to-one mux 36, the imaginary data is inserted before the real data. This operation causes the real data and imaginary data to be skewed by one clock cycle. The two-to-one mux 56 and the two-to-one mux 58 insert the real data before the imaginary data so that the one clock cycle skew does not grow. A ten-bit register 64 shown in Figure 3 is placed after the two-to-one mux 62 in order to provide a one clock cycle delay to remove this skew.

A numerically controlled oscillator 66 has two outputs, an eight-bit sine wave output and an eight-bit cosine wave output. The phase accumulator of the numerically controlled oscillator 66 is sixteen bits. For channel 3, the phase increment is 17,582, and the output frequency is 23.098348 MHZ. For channel 4, the phase increment is 13,015, and the output frequency is 17.098453 MHZ. The frequency of the pilot (f_{out}) is determined according to the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1) \qquad (2)$$

where f_1 represents the symbol frequency, where the factor $8f_1$ results from the three interpolations-by-two, where the factor $f_1/4$ results from the 2's complement circuit 26, and where the factor f_2 represents the output of the numerically controlled oscillator 66. Accordingly, the pilot frequency for channel 3 is 60.308995 MHZ, and the pilot frequency for channel 4 is 66.308890 MHZ. Because the phase accumulator is only sixteen bits, the pilot frequencies are slightly below the ideal pilot frequencies. The channel select signal controls the NCO frequency. For example, a logic level zero selects channel 3, and a logic level one selects channel 4.

The complex data is modulated by the quadrature output of the numerically controlled oscillator 66. A multiplier 68 multiplies the ten-bit real component of the symbol data from the two-to-one mux 60 by the eight-bit cosine output of the numerically controlled oscillator 66. A multiplier 70 multiplies the ten-bit imaginary component of the symbol data from the two-to-one mux 62 and the ten-bit register 64 by the eight-bit sine output of the numerically controlled oscillator 66. A summer 72 subtracts the output of the multiplier 70 from the output of the

multiplier 68 which is equivalent to shifting the frequency spectrum of the complex data and its conjugate by the NCO frequency and summing the two together. This operation is shown by the following equation:

$$x_n = \frac{(r_n + ji_n)e^{jn\omega nco} + (r_n - ji_n)e^{-jn\omega nco}}{2} \qquad (3)$$

$$= r_n\cos(\omega_{nco}n) - i_n\sin(\omega_{nco}n)$$

The modulated data has a center frequency equal to the output frequency of the numerically controlled oscillator 66. The pilot frequency is 2.69 MHZ above the output frequency of the numerically controlled oscillator 66. Figure 8a shows the frequency spectrum of the modulated data on channel 3, and Figure 8b shows the frequency spectrum of the modulated data on channel 4.

The output of the summer 72 is attenuated by an attenuator 74 and converted to an analog signal by a digital-to-analog converter 76. The attenuation due to the sinx/x roll-off of the digital-to-analog converter 76 can be greater for channel 4 than it is for channel 3. Also, the DC gain of the sinx/x correction filter 20 can be slightly

higher for channel 3 than for channel 4. Therefore, if these conditions exist, the signal power after the digital-to-analog converter 76 is greater (for example, by about 3 db) for channel 3 than for channel 4. Accordingly, the attenuator 74 is provided to apply a suitable attenuation (such as 23/32) to make the output power for channel 3 and channel 4 substantially equal. Thus, when the channel select signal indicates channel 3, the attenuation applied by the attenuator 74 is made active and, when the channel select signal indicates channel 4, the attenuation applied by the attenuator 74 is made inactive. The attenuator 74 may also be arranged to round the data to nine bits and to limit the data to eight bits regardless of channel.

The digital-to-analog converter 76 has an eight bit resolution and a differential output. This differential output is converted to a single ended output through a 1:1 transformer 78 of the output circuit 16 as shown in Figure 10. The side of the 1:1 transformer 78 coupled to the digital-to-analog converter 76 has a grounded center tap and the output impedance of the digital-to-analog converter 76 is set by placing a 75Ω resistor across the output of the digital-to-analog converter 76.

~~120~~

Figure 9a shows the frequency spectrum at the output of the digital-to-analog converter 76 for channel 3, and Figure 9b shows the frequency spectrum at the output of the digital-to-analog converter 76 for channel 4. The frequency of the numerically controlled oscillator 66 may be chosen so that the first image of the modulated data above the Nyquist frequency falls substantially in either channel 3 or channel 4. The sinx/x correction imposed by the sinx/x correction filter 20 may be also designed to correct this image. The signal-to-noise ratio of the first image is also good enough to be received by the receiver. Therefore, the first image may be used as the modulated output and the fundamental output of the digital-to-analog converter 76 can be ignored.

Accordingly, a bandpass filter 80 of the output circuit 16 as shown in Figure 10 is arranged to remove the modulated data below channels 3 and 4 and all images above channels 3 and 4. An attenuator 82 lowers the signal power to a nominal 0 dbmv.

Certain modifications and alternatives of the present invention have been discussed above. Other modifications and alternatives will occur to those practicing in

the art of the present invention. Accordingly, the description of the present invention is to be construed as illustrative only and is for the purpose of teaching those skilled in the art the best mode of carrying out the invention. The details may be varied substantially without departing from the spirit of the invention, and the exclusive use of all modifications which are within the scope of the appended claims is reserved.

WHAT IS CLAIMED IS:

1. A system for supplying a compressed digital signal to a digital television receiver comprising:

a source of compressed baseband signal;

an encoder that encodes the baseband signal in a digital format, wherein the baseband signal has a frequency;

a first frequency shifter that shifts the frequency of the encoded baseband signal;

a first interpolator that interpolates an output of the first frequency shifter;

a second interpolator that interpolates an output of the first interpolator;

a third interpolator that interpolates an output of the second interpolator; and,

a second frequency shifter that shifts a frequency of an output of the third interpolator.

2. The system of claim 1 wherein the encoder provides the baseband signal at a frequency f_1, and wherein the first frequency shifter shifts the baseband signal by $f_1/4$.

3. The system of claim 2 wherein the first interpolator interpolates by two, wherein the second interpolator interpolates by two, and wherein the third interpolator interpolates by two.

4. The system of claim 2 wherein the first, second, and third interpolators collectively interpolate by eight.

5. The system of claim 4 wherein the first and second frequency shifters and the first, second, and third interpolators effectively operate in accordance with the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the first, second, and third interpolators, where $f_1/4$ results from the first frequency shifter, and where f_2 is supplied by the second frequency shifter.

6. The system of claim 1 further comprising a low pass filter coupled between the encoder and the first interpolator, wherein the encoder provides the baseband signal at a frequency f_1, and wherein the low pass filter has a response that rolls off at about $f_1/4$.

7. The system of claim 6 wherein the first frequency shifter shifts the baseband signal by $f_1/4$.

8. The system of claim 7 wherein the first, second, and third interpolators collectively interpolate by eight.

9. The system of claim 8 wherein the first and second frequency shifters and the first, second, and third interpolators effectively operate in accordance with the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the first, second, and third interpolators, where $f_1/4$ results from the first frequency

shifter, and where f_2 is supplied by the second frequency shifter.

10. The system of claim 1 further comprising a sinx/x correction coupled between the encoder and the first interpolator, wherein the sinx/x correction provides compensation for the sinx/x roll off created by processing in the system.

11. The system of claim 10 further comprising a low pass filter coupled between the encoder and the first interpolator, wherein the encoder provides the baseband signal at a frequency f_1, and wherein the low pass filter has a response that rolls off at about $f_1/4$.

12. The system of claim 10 wherein the encoder provides the baseband signal at a frequency f_1, and wherein the first frequency shifter shifts the baseband signal by $f_1/4$.

13. The system of claim 12 wherein the first, second, and third interpolators collectively interpolate by eight.

14. The system of claim 13 wherein the first and second frequency shifters and the first, second, and third interpolators effectively operate in accordance with the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the first, second, and third interpolators, where $f_1/4$ results from the first frequency shifter, and where f_2 is supplied by the second frequency shifter.

15. The system of claim 1 wherein the first interpolator interpolates by two, wherein the second interpolator interpolates by two, and wherein the third interpolator interpolates by two.

16. A method of supplying a compressed digital signal to a digital television receiver comprising:

encoding a baseband signal in a digital format, wherein the baseband signal has a symbol frequency;

increasing the symbol frequency of the encoded baseband signal to produce an intermediate frequency signal, wherein the increasing of the symbol frequency of the encoded baseband signal includes three interpolations of the encoded baseband signal; and,

modulating the intermediate frequency signal with a channel selection frequency signal so as to produce an output signal having the frequency of a selected channel.

17. The method of claim 16 wherein the baseband signal has a symbol frequency f_1, and wherein the increasing of the symbol frequency of the encoded baseband signal comprises shifting the symbol frequency of the baseband signal by $f_1/4$.

18. The method of claim 16 wherein the three interpolations are each an interpolation by two.

19. The method of claim 16 wherein the baseband signal has a symbol frequency f_1, and wherein the output signal has a frequency f_{out} given by the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the three interpolations, where $f_1/4$ results from a frequency shifting of the encoded baseband signal, and where f_2 is the frequency of the channel selection frequency signal.

20. The method of claim 16 further comprising low pass filtering the encoded baseband signal, wherein the baseband signal has a symbol frequency f_1, and wherein the low pass filtering has a response that rolls off at about $f_1/4$.

21. The method of claim 20 wherein the increasing of the symbol frequency of the encoded baseband signal comprises shifting the symbol frequency of the baseband signal by $f_1/4$.

22. The method of claim 20 wherein the three interpolations are each an interpolation by two.

23. The method of claim 20 wherein the output signal has a frequency f_{out} given by the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the three interpolations, where $f_1/4$ results from a frequency shifting of the encoded baseband signal, and where f_2 is the frequency of the channel selection frequency signal.

24. The method of claim 16 further comprising sinx/x correcting the encoded baseband signal, wherein the sinx/x correction compensates for the sinx/x roll off created by processing downstream of the sinx/x correction.

25. The method of claim 24 further comprising low pass filtering the encoded baseband signal, wherein the baseband signal has a symbol frequency f_1, and wherein the low pass filtering has a response that rolls off at about $f_1/4$.

26. The method of claim 24 wherein the baseband signal has a symbol frequency f_1, and wherein the increasing of the symbol frequency of the encoded baseband signal comprises shifting the symbol frequency of the baseband signal by $f_1/4$.

27. The method of claim 24 wherein the three interpolations are each an interpolation by two.

28. The method of claim 24 wherein the baseband signal has a symbol frequency f_1, and wherein the output signal has a frequency f_{out} given by the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the three interpolations, where $f_1/4$ results from a frequency shifting of the encoded baseband signal, and where f_2 is the frequency of the channel selection frequency signal.

29. The method of claim 16 wherein the output signal contains modulated data below the selected channel and an image at the selected channel, and wherein the method comprises filtering of the output signal to remove the modulated data and to pass the image.

30. An electrical signal produced by (i) encoding a signal in a digital television format, wherein the signal has a symbol frequency f_1, (ii) shifting the frequency of the encoded signal by $f_1/4$, (iii) increasing the frequency of the shifted encoded signal to produce an intermediate signal, and (iv) modulating the intermediate frequency signal with a channel selection frequency signal so as to produce the electrical signal at the frequency of a selected channel.

31. The electrical signal of claim 30 wherein the frequency of the shifted encoded signal is increased by interpolation, and wherein the interpolation comprises three interpolations by two.

32. The electrical signal of claim 30 having a frequency f_{out} given by the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the increasing of the frequency of the shifted encoded signal, where $f_1/4$ results from the frequency shifting, and where f_2 is the frequency of the channel selection frequency signal.

33. The electrical signal of claim 30 wherein the encoded signal is low pass filtered, and wherein the low pass filtering has a response that rolls off at about $f_1/4$.

34. The electrical signal of claim 33 wherein the frequency of the shifted encoded signal is increased by interpolation, and wherein the interpolation comprises three interpolations by two.

35. The electrical signal of claim 33 having a frequency f_{out} given by the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the increasing of the frequency of the shifted encoded signal, where $f_1/4$ results from the frequency shifting, and where f_2 is the frequency of the channel selection frequency signal.

36. The electrical signal of claim 30 wherein the encoded signal is sinx/x corrected, and wherein the sinx/x correction compensates for the sinx/x roll off created by processing of the electrical signal downstream of the sinx/x correction.

37. The electrical signal of claim 36 wherein the encoded signal is low pass filtered, and wherein the low pass filtering has a response that rolls off at about $f_1/4$.

38. The electrical signal of claim 36 wherein the frequency of the shifted encoded signal is increased by interpolation, and wherein the interpolation comprises three interpolations by two.

39. The electrical signal of claim 36 having a frequency f_{out} given by the following equation:

$$f_{out} = 8f_1 - (f_2 + \frac{1}{4}f_1)$$

where $8f_1$ results from the increasing of the frequency of the shifted encoded signal, where $f_1/4$ results from the frequency shifting, and where f_2 is the frequency of the channel selection frequency signal.

40. The electrical signal of claim 30 further produced by filtering so as remove data below the selected channel and so as to leave an image of the data as the electrical signal.

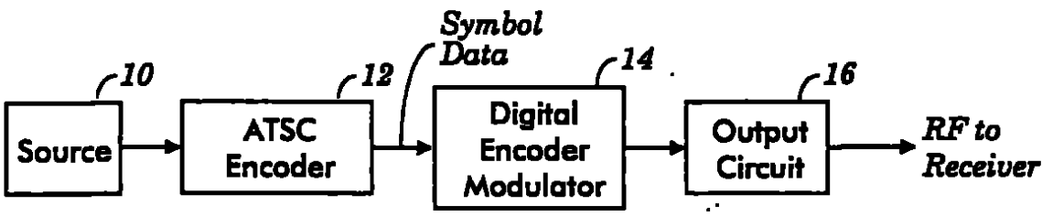

Symbol Data
10
12
14
16
Source
ATSC Encoder
Digital Encoder Modulator
Output Circuit
RF to Receiver

Figure 1

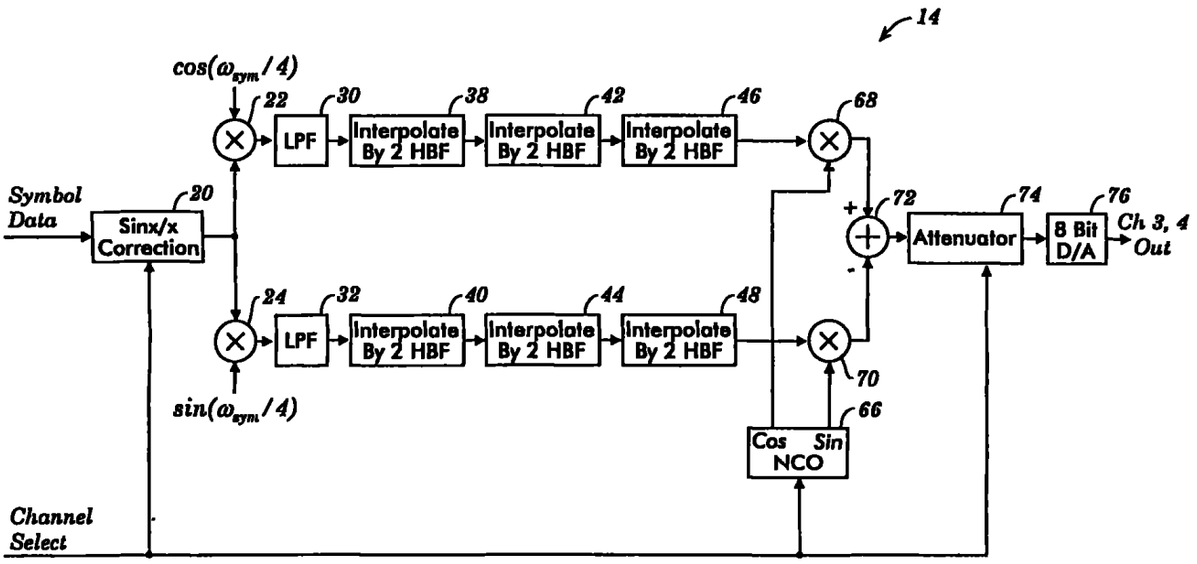

14
cos(ω_sym / 4)
22
30
38
42
46
68
LPF
Interpolate By 2 HBF
Interpolate By 2 HBF
Interpolate By 2 HBF
Symbol Data
20
Sinx/x Correction
72
74
76
Attenuator
8 Bit D/A
Ch 3, 4 Out
24
32
40
44
48
LPF
Interpolate By 2 HBF
Interpolate By 2 HBF
Interpolate By 2 HBF
70
sin(ω_sym / 4)
66
Cos Sin NCO
Channel Select

Figure 2

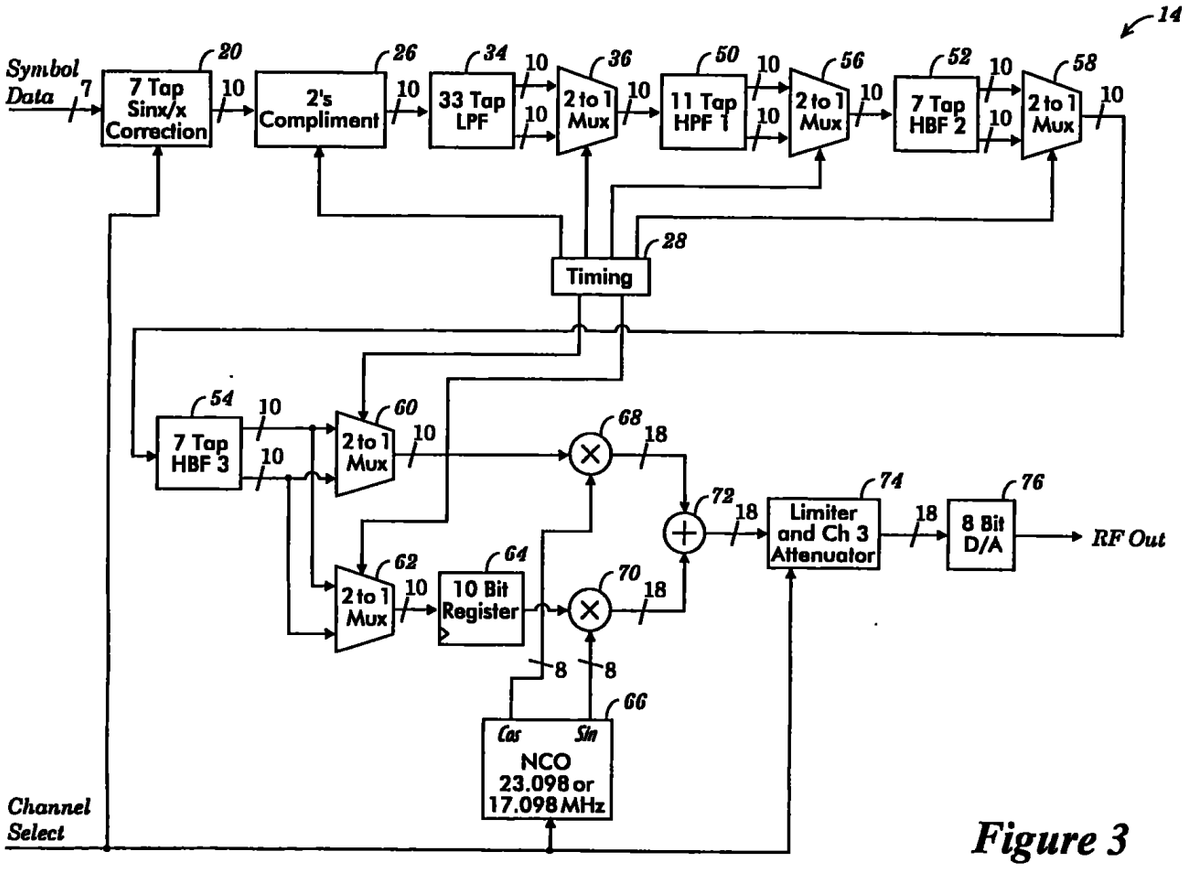

14
Symbol Data
7
20
7 Tap Sinx/x Correction
10
26
2's Compliment
34
33 Tap LPF
36
2 to 1 Mux
50
11 Tap HPF 1
56
2 to 1 Mux
52
7 Tap HBF 2
58
2 to 1 Mux
28
Timing
54
7 Tap HBF 3
60
2 to 1 Mux
62
2 to 1 Mux
64
10 Bit Register
68
70
72
74
Limiter and Ch 3 Attenuator
76
8 Bit D/A
RF Out
18
8
66
Cos
Sin
NCO 23.098 or 17.098 MHz
Channel Select

Figure 3

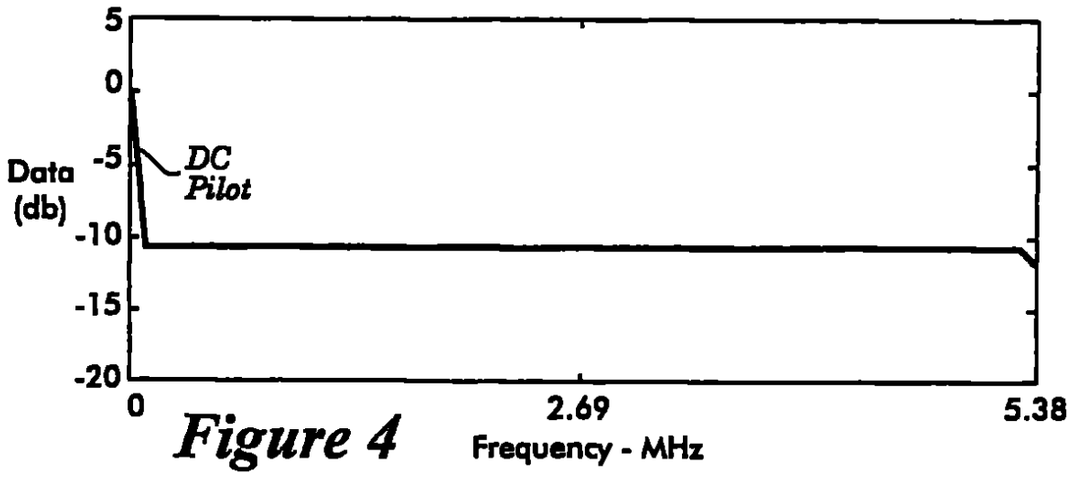


Figure 4

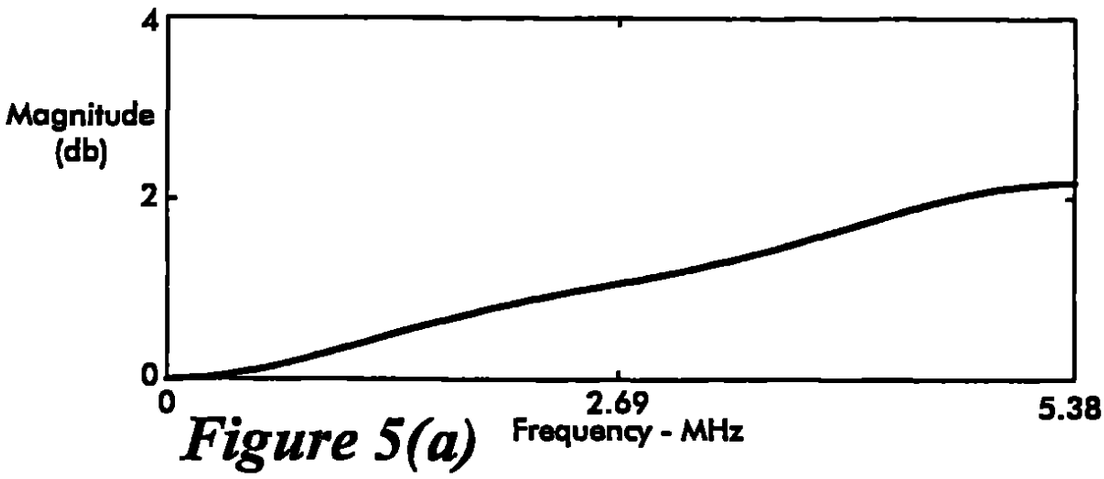


Figure 5(a)

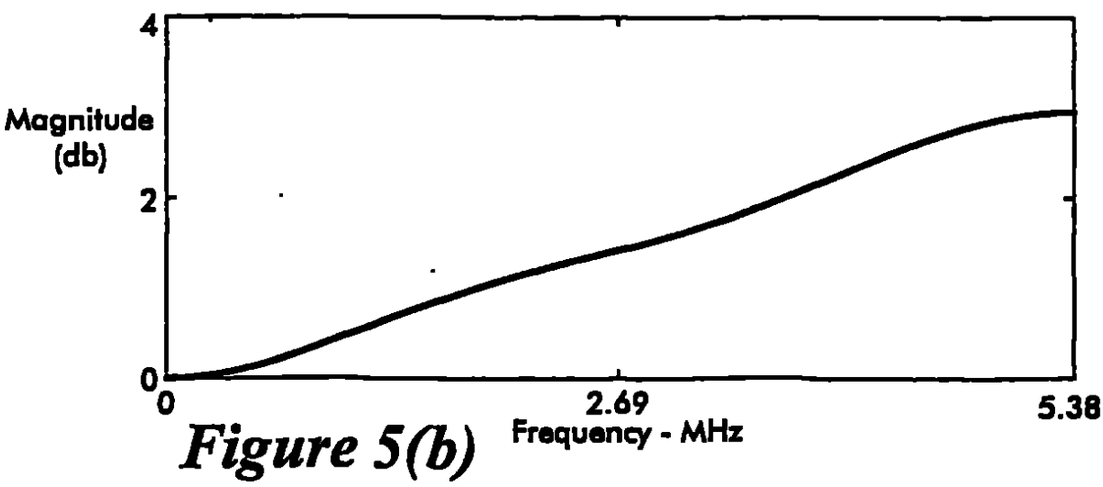


Figure 5(b)

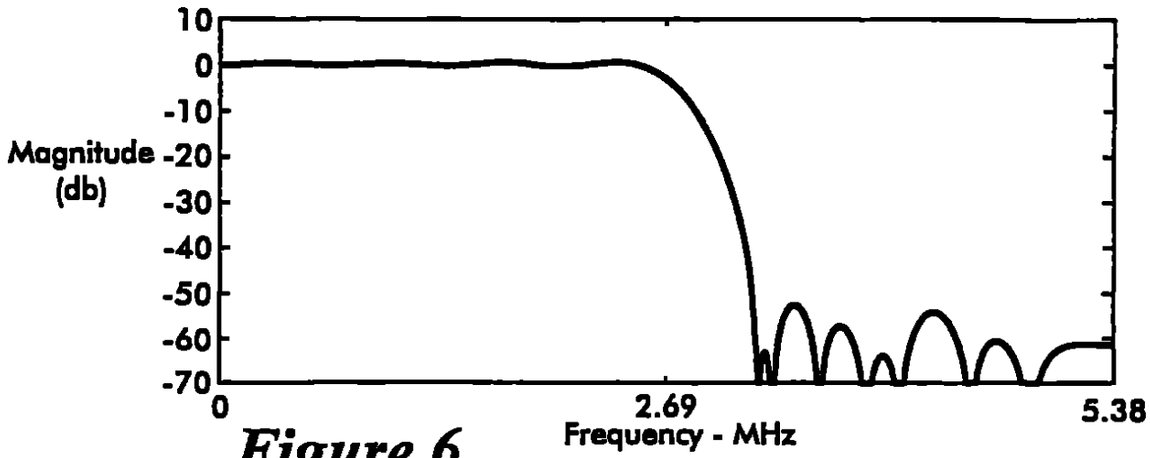


Figure 6

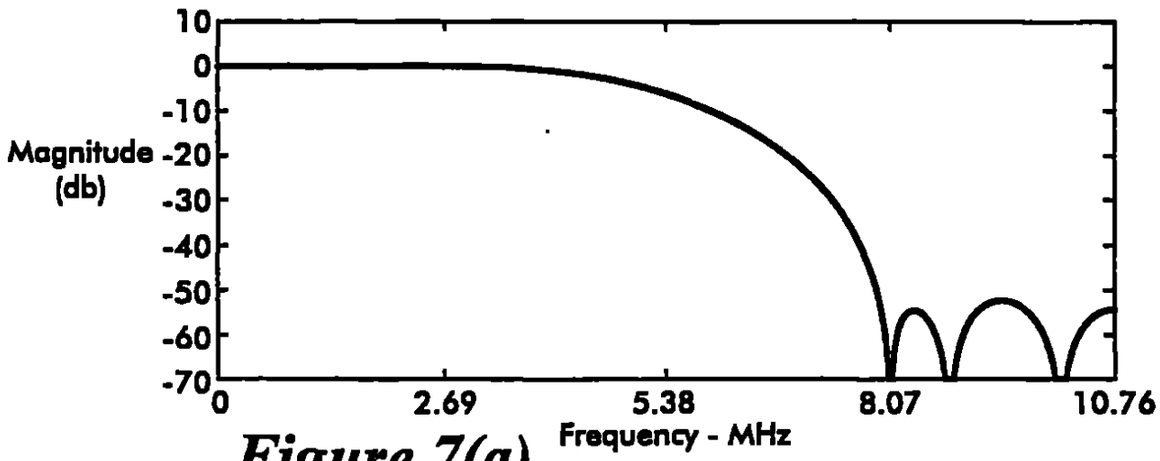


Figure 7(a)

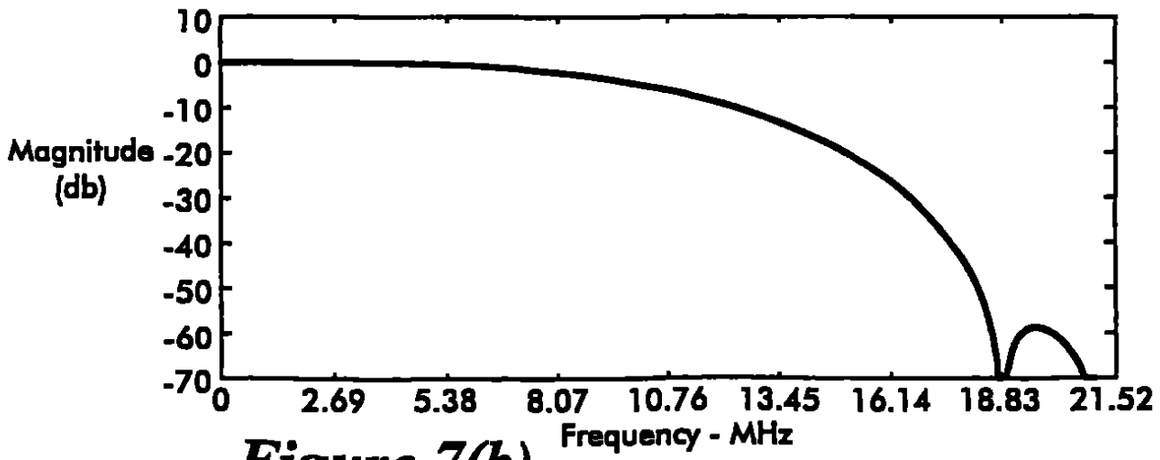


Figure 7(b)

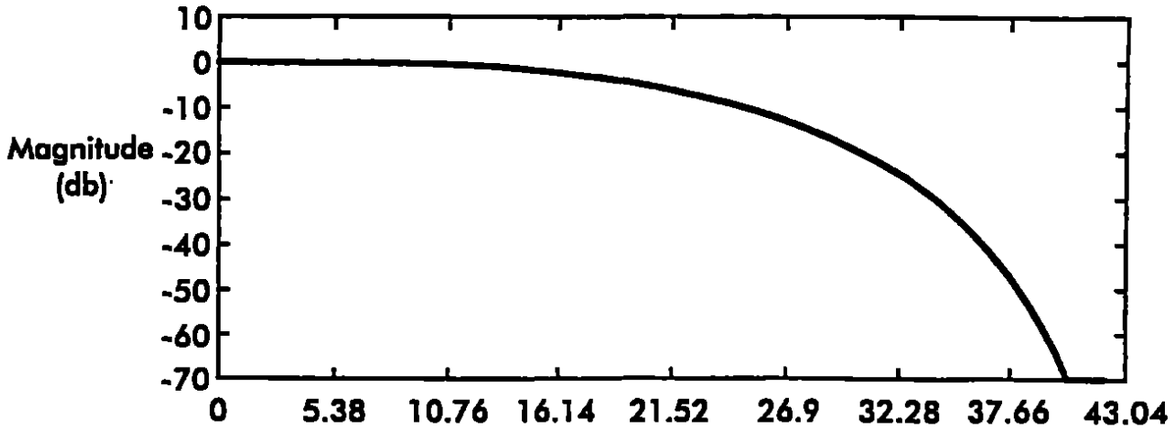


Figure 7(c)

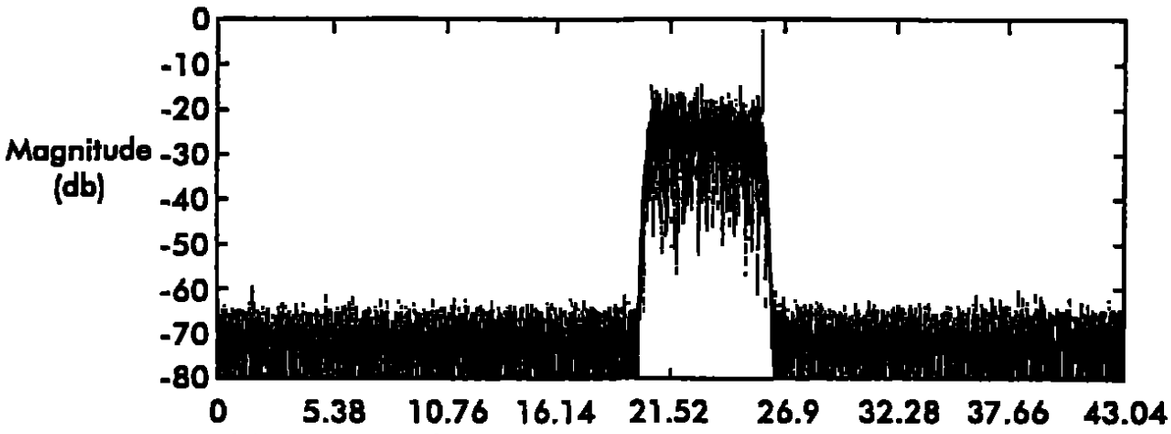


Figure 8(a)

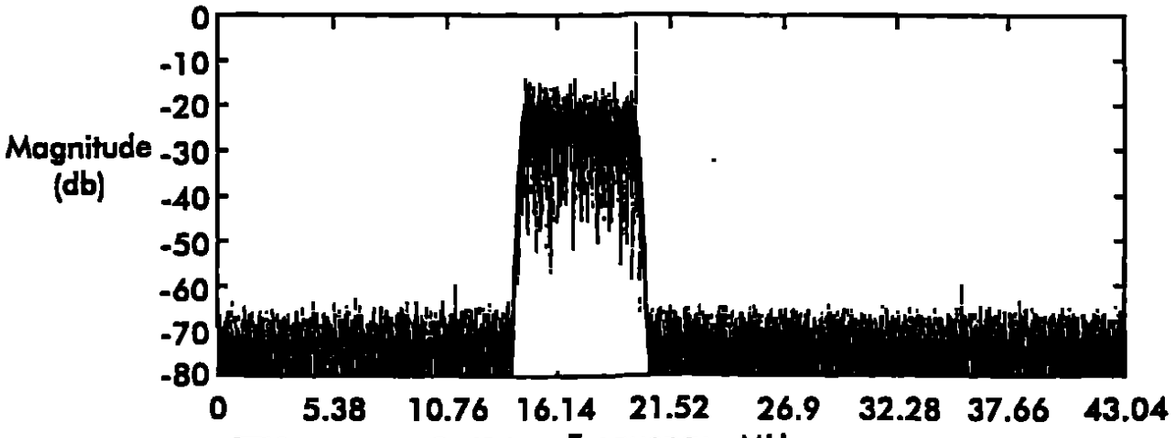


Figure 8(b)

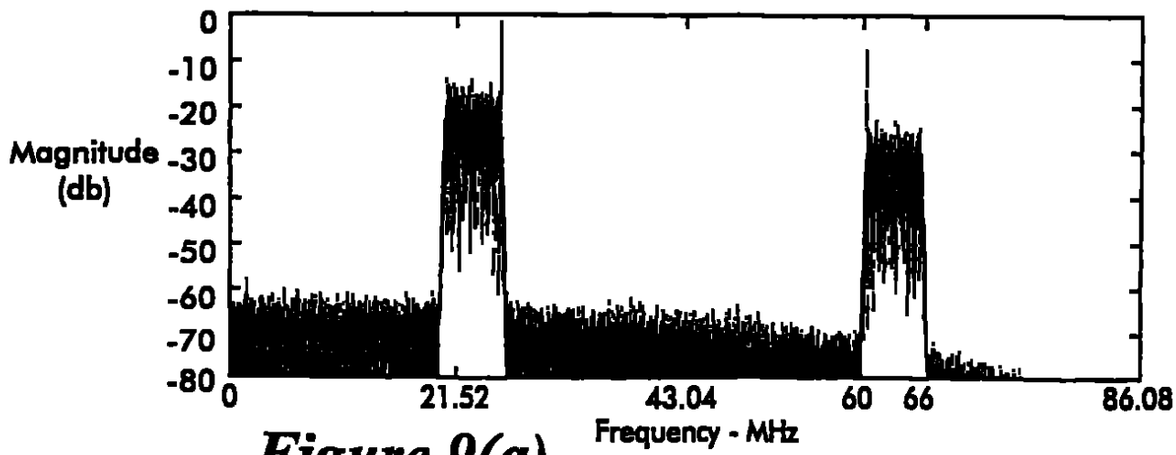


Figure 9(a)

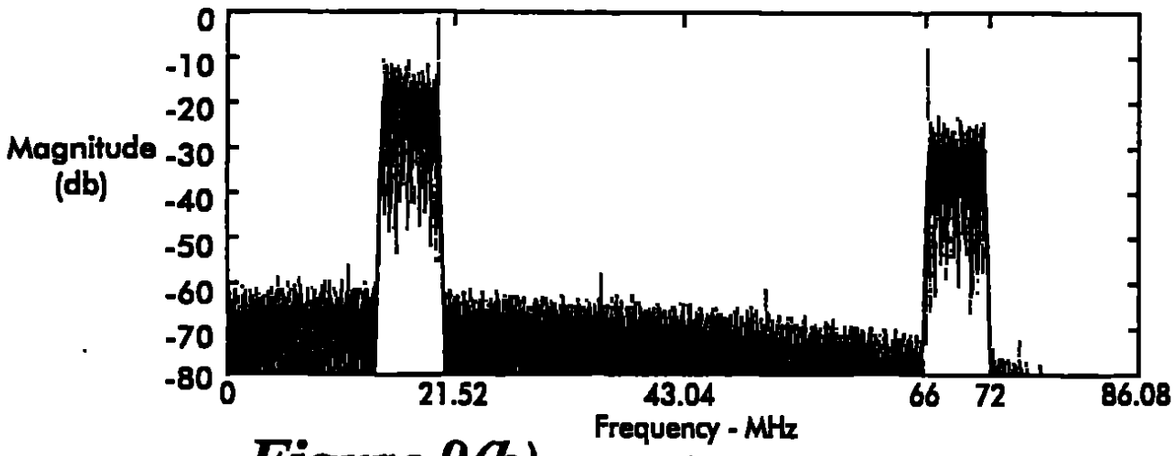


Figure 9(b)

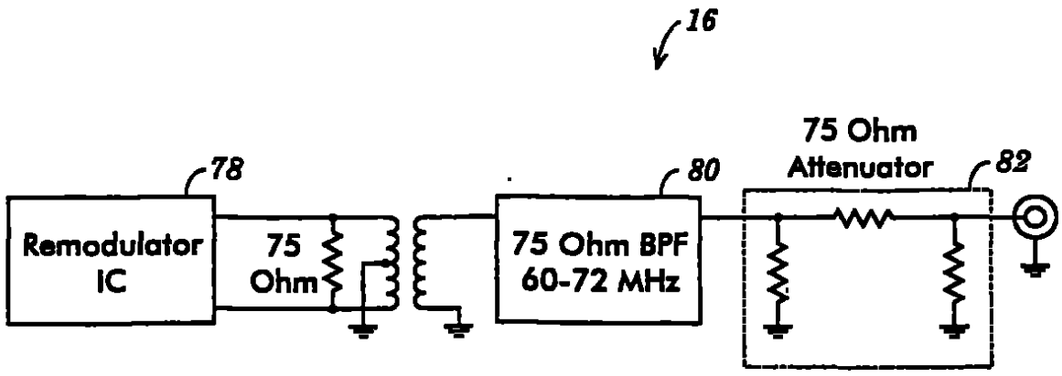


Figure 10

141 RECEIVED JUN 2 2002 SCHIFF HARDIN & WAITE IIS PATENT OFFICE 141

PATENT COOPERATION TREATY

From the
INTERNATIONAL PRELIMINARY EXAMINING AUTHORITY

PCT

To:
Joike, Trevor B.
SCHIFF HARDIN & WAITE
6600 Sears Tower
Chicago, Illinois 60606
ETATS-UNIS D'AMERIQUE

WRITTEN OPINION

(PCT Rule 66)

Date of mailing *(day/month/year)*	21/06/2002
Applicant's or agent's file reference: FP010516WO2947136818	REPLY DUE: within 1 / 00 months/days from the above date of mailing

International application No.	International filing date *(day/month/year)*	Priority date *(day/month/year)*
PCT/US 01/ 23211	23/07/2001	04/08/2000

International Patent Classification (IPC) or both national classification and IPC
H04N7/24

Applicant
ZENITH ELECTRONICS CORPORATION

1. This written opinion is the first drawn up by this International Preliminary Examining Authority.
2. This opinion contains indications relating to the following items:
   - I [X] Basis of the opinion
   - II [ ] Priority
   - III [ ] Non-establishment of opinion with regard to novelty, inventive step and industrial applicability
   - IV [ ] Lack of unity of invention
   - V [X] Reasoned statement under Rule 66.2(a)(ii) with regard to novelty, inventive step or industrial applicability; citations and explanations supporting such statement
   - VI [ ] Certain documents cited
   - VII [ ] Certain defects in the international application
   - VIII [ ] Certain observations on the international application
3. The applicant is hereby invited to reply to this opinion.
   - **When?** See the time limit indicated above. The applicant may, before the expiration of that time limit, request this Authority to grant an extension, see Rule 66.2(d).
   - **How?** By submitting a written reply, accompanied, where appropriate, by amendments, according to Rule 66.3. For the form and the language of the amendments, see Rules 66.8 and 66.9.
   - **Also** For an additional opportunity to submit amendments, see Rule 66.4. For the examiner's obligation to consider amendments and/or arguments, see Rule 66.4*bis*. For an informal communication with the examiner, see Rule 66.6.

   If no reply is filed, the international preliminary examination report will be established on the basis of this opinion.
4. The final date by which the international preliminary examination report must be established according to Rule 69.2 is: 04/12/2002

Name and mailing address of the IPEA/
European Patent Office
D-80298 Munich
Tel. (+49-89) 2399-0, Tx: 523656 epmu d
Fax: (+49-89) 2399-4465

Authorized officer
Examiner

Formalities officer
(incl. extension of time limits)
Tel. (+49-89) 2399 2828

142

I. Basis of the opinion

1. The basis of this written opinion is the application as originally filed.

V. Reasoned statement under Rule 66.2(a)(ii) with regard to novelty, inventive step or industrial applicability

1. In light of the documents cited in the international search report, it is considered that the invention as defined in at least some of the claims does not appear to meet the criteria mentioned in Article 33(1) PCT, i.e. does not appear to be novel and/or to involve an inventive step (see international search report, in particular the documents cited X and/or Y and corresponding claims references).

2. If amendments are filed, the applicant should comply with the requirements of Rule 66.8 PCT and indicate the basis of the amendments in the documents of the application as originally filed (Article 34 (2) (b) PCT) otherwise these amendments may not be taken into consideration for the establishment of the international preliminary examination report. The attention of the applicant is drawn to the fact that if the application contains an unnecessary plurality of independent claims, no examination of any of the claims will be carried out.

NB: Should the applicant decide to request detailed substantive examination, then an international preliminary examination report will normally be established directly. Exceptionally the examiner may draw up a second written opinion, should this be explicitly requested.

PATENT COOPERATION TREATY

143

From the
INTERNATIONAL PRELIMINARY EXAMINING AUTHORITY

PCT

143

To:
Joike, Trevor B.
SCHIFF HARDIN & WAITE
6600 Sears Tower
Chicago, Illinois 60606
ETATS-UNIS D'AMERIQUE

NOTIFICATION OF TRANSMITTAL OF INTERNATIONAL PRELIMINARY EXAMINATION REPORT

(PCT Rule 71.1)

Date of mailing *(day/month/year)*	12/08/2002
Applicant's or agent's file reference: FP010516W02947136818	IMPORTANT NOTIFICATION

International application No.	International filing date *(day/month/year)*	Priority date *(day/month/year)*
PCT/US 01/ 23211	23/07/2001	04/08/2000

Applicant

ZENITH ELECTRONICS CORPORATION

1. The applicant is hereby notified that this International Preliminary Examining Authority transmits herewith the international preliminary examination report and its annexes, if any, established on the international application.

2. A copy of the report and its annexes, if any, is being transmitted to the International Bureau for communication to all the elected Offices.

3. Where required by any of the elected Offices, the International Bureau will prepare an English translation of the report (but not of any annexes) and will transmit such translation to those Offices.

4. **REMINDER**

The applicant must enter the national phase before each elected Office by performing certain acts (filing translations and paying national fees) within 30 months from the priority date (or later in some Offices)(Article 39(1))(see also the reminder sent by the International Bureau with Form PCT/IB/301).

Where a translation of the international application must be furnished to an elected Office, that translation must contain a translation of any annexes to the international preliminary examination report. It is the applicant's responsibility to prepare and furnish such translation directly to each elected Office concerned.

For further details on the applicable time limits and requirements of the elected Offices, see Volume II of the PCT Applicant's Guide.

For the purpose of deciding whether the claimed invention is patentable or not, the elected Offices may apply criteria additional to or different from the criteria on which the international preliminary examination report is based (see Articles 27 (5), 33 (5)). Additional criteria may include e.g. exemptions from patentability and the requirements of enabling disclosure and of clarity and support of claims.

RECEIVED
[illegible] 2002
SCHIFF HARDIN & WAITE
PATENT OFFICE

Name and mailing address of the IPEA/	Authorized officer
European Patent Office D-80298 Munich Tel. (+49-89) 2399-0, Tx: 523656 epmu d Fax: (+49-89) 2399-4465	BIERMAIER R M Tel. (+49-89) 2399 2828

Form PCT/IPEA/416 (28/06/2002) P20473

144

146

PATENT COOPERATION TREATY

PCT

INTERNATIONAL PRELIMINARY EXAMINATION REPORT

(PCT Article 36 and Rule 70)

Applicant's or agent's file reference	FOR FURTHER ACTION	See Notification of Transmittal of International Preliminary Examination Report (Form PCT/IPEA/416)
FP010516W02947136818		
International application No.	International filing date *(day/month/year)*	Priority date *(day/month/year)*
PCT/US 01/ 23211	23/07/2001	04/08/2000
International Patent Classification (IPC) or national classification and IPC		
H04N7/24		
Applicant		
ZENITH ELECTRONICS CORPORATION		

1. This international preliminary examination report has been prepared by this International Preliminary Examining Authority and is transmitted to the applicant according to Article 36.

2. This REPORT consists of a total of 2 sheets, including this cover sheet.

☐ This report is also accompanied by ANNEXES, i.e., sheets of the description, claims and/or drawings which have been amended and are the basis for this report and/or sheets containing rectifications made before this Authority (see Rule 70.16 and Section 607 of the Administrative Instructions under the PCT).

These annexes consists of a total of _____ sheets.

3. This report contains indications relating to the following items:

- I ☒ Basis of the report
- II ☐ Priority
- III ☐ Non-establishment of opinion with regard to novelty, inventive step and industrial applicability
- IV ☐ Lack of unity of invention
- V ☒ Reasoned statement under Article 35(2) with regard to novelty, inventive step or industrial applicability; citations and explanations supporting such statement
- VI ☐ Certain documents cited
- VII ☐ Certain defects in the international application
- VIII ☐ Certain observations on the international application

Date of submission of the demand	Date of completion of this report
11/02/2002	07/08/2002
Name and mailing address of the IPEA/	Authorized officer
European Patent Office D-80298 Munich Tel. (+49-89) 2399-0, Tx: 523656 epmu d Fax: (+49-89) 2399-4465	CHAIX DE LAVARENE C P M Tel. (+49-89) 2399 2828

Form PCT/IPEA/409 (cover sheet) (July 1998)

I. Basis of the report

The basis of this international preliminary examination is the application as originally filed.

V. Reasoned statement under Rule 66.2(a)(ii) with regard to novelty, inventive step or industrial applicability

In light of the documents cited in the international search report, it is considered that the invention as defined in at least some of the claims does not appear to meet the criteria mentioned in Article 33(1) PCT, i.e. does not appear to be novel and/or to involve an inventive step (see international search report, in particular the documents cited X and/or Y and corresponding claim references).

PATENT COOPERATION TREATY

146

From the INTERNATIONAL SEARCHING AUTHORITY

PCT

NOTIFICATION OF TRANSMITTAL OF THE INTERNATIONAL SEARCH REPORT OR THE DECLARATION

(PCT Rule 44.1)

To:
MARSHALL O'TOOLE GERSTEIN
MURRAY & BORUN
Attn. Joike, Trevor B.
6300 Sears Tower
233 South Wacker Drive
Chicago, Illinois 60606
UNITED STATES OF AMERICA

Date of mailing *(day/month/year)*	20/03/2002
Applicant's or agent's file reference: 29471/36818	FOR FURTHER ACTION See paragraphs 1 and 4 below
International application No.: PCT/US 01/23211	International filing date *(day/month/year)*: 23/07/2001
Applicant: ZENITH ELECTRONICS CORPORATION	

1. [X] The applicant is hereby notified that the International Search Report has been established and is transmitted herewith.

   Filing of amendments and statement under Article 19:
   The applicant is entitled, if he so wishes, to amend the claims of the International Application (see Rule 46):

   When? The time limit for filing such amendments is normally 2 months from the date of transmittal of the International Search Report; however, for more details, see the notes on the accompanying sheet.

   Where? Directly to the International Bureau of WIPO
   34, chemin des Colombettes
   1211 Geneva 20, Switzerland
   Fascimile No.: (41-22) 740.14.35

   For more detailed instructions, see the notes on the accompanying sheet.

2. [ ] The applicant is hereby notified that no International Search Report will be established and that the declaration under Article 17(2)(a) to that effect is transmitted herewith.

3. [ ] With regard to the protest against payment of (an) additional fee(s) under Rule 40.2, the applicant is notified that:

   [ ] the protest together with the decision thereon has been transmitted to the International Bureau together with the applicant's request to forward the texts of both the protest and the decision thereon to the designated Offices.

   [ ] no decision has been made yet on the protest; the applicant will be notified as soon as a decision is made.

4. **Further action(s):** The applicant is reminded of the following:

   Shortly after 18 months from the priority date, the international application will be published by the International Bureau. If the applicant wishes to avoid or postpone publication, a notice of withdrawal of the international application, or of the priority claim, must reach the International Bureau as provided in Rules 90*bis*.1 and 90*bis*.3, respectively, before the completion of the technical preparations for international publication.

   Within 19 months from the priority date, a demand for international preliminary examination must be filed if the applicant wishes to postpone the entry into the national phase until 30 months from the priority date (in some Offices even later).

   Within 20 months from the priority date, the applicant must perform the prescribed acts for entry into the national phase before all designated Offices which have not been elected in the demand or in a later election within 19 months from the priority date or could not be elected because they are not bound by Chapter II.

Name and mailing address of the International Searching Authority	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL-2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Jolanda Offerman-Hazeleger

Form PCT/ISA/220 (July 1998)

NOTES TO FORM PCT/ISA/220

147

These Notes are intended to give the basic instructions concerning the filing of amendments under article 19. The Notes are based on the requirements of the Patent Cooperation Treaty, the Regulations and the Administrative Instructions under that Treaty. In case of discrepancy between these Notes and those requirements, the latter are applicable. For more detailed information, see also the PCT Applicant's Guide, a publication of WIPO.

In these Notes, "Article", "Rule", and "Section" refer to the provisions of the PCT, the PCT Regulations and the PCT Administrative Instructions respectively.

INSTRUCTIONS CONCERNING AMENDMENTS UNDER ARTICLE 19

The applicant has, after having received the international search report, one opportunity to amend the claims of the international application. It should however be emphasized that, since all parts of the international application (claims, description and drawings) may be amended during the international preliminary examination procedure, there is usually no need to file amendments of the claims under Article 19 except where, e.g. the applicant wants the latter to be published for the purposes of provisional protection or has another reason for amending the claims before international publication. Furthermore, it should be emphasized that provisional protection is available in some States only.

What parts of the international application may be amended?

Under Article 19, only the claims may be amended.

During the international phase, the claims may also be amended (or further amended) under Article 34 before the International Preliminary Examining Authority. The description and drawings may only be amended under Article 34 before the International Examining Authority.

Upon entry into the national phase, all parts of the international application may be amended under Article 28 or, where applicable, Article 41.

When?

Within 2 months from the date of transmittal of the international search report or 16 months from the priority date, whichever time limit expires later. It should be noted, however, that the amendments will be considered as having been received on time if they are received by the International Bureau after the expiration of the applicable time limit but before the completion of the technical preparations for international publication (Rule 46.1).

Where not to file the amendments?

The amendments may only be filed with the International Bureau and not with the receiving Office or the International Searching Authority (Rule 46.2).

Where a demand for international preliminary examination has been/is filed, see below.

How?

Either by cancelling one or more entire claims, by adding one or more new claims or by amending the text of one or more of the claims as filed.

A replacement sheet must be submitted for each sheet of the claims which, on account of an amendment or amendments, differs from the sheet originally filed.

All the claims appearing on a replacement sheet must be numbered in Arabic numerals. Where a claim is cancelled, no renumbering of the other claims is required. In all cases where claims are renumbered, they must be renumbered consecutively (Administrative Instructions, Section 205(b)).

The amendments must be made in the language in which the international application is to be published.

What documents must/may accompany the amendments?

Letter (Section 205(b)):

The amendments must be submitted with a letter.

The letter will not be published with the international application and the amended claims. It should not be confused with the "Statement under Article 19(1)" (see below, under "Statement under Article 19(1)").

The letter must be in English or French, at the choice of the applicant. However, if the language of the international application is English, the letter must be in English; if the language of the international application is French, the letter must be in French.

BNSDOCID <XISA220NOENP4 I >

140
143

The letter must indicate the differences between the claims as filed and the claims as amended. It must, in particular, indicate, in connection with each claim appearing in the international application (it being understood that identical indications concerning several claims may be grouped), whether

(i) the claim is unchanged;

(ii) the claim is cancelled;

(iii) the claim is new;

(iv) the claim replaces one or more claims as filed;

(v) the claim is the result of the division of a claim as filed.

The following examples illustrate the manner in which amendments must be explained in the accompanying letter:

1. [Where originally there were 48 claims and after amendment of some claims there are 51]:
"Claims 1 to 29, 31, 32, 34, 35, 37 to 48 replaced by amended claims bearing the same numbers; claims 30, 33 and 36 unchanged; new claims 49 to 51 added."

2. [Where originally there were 15 claims and after amendment of all claims there are 11]:
"Claims 1 to 15 replaced by amended claims 1 to 11."

3. [Where originally there were 14 claims and the amendments consist in cancelling some claims and in adding new claims]:
"Claims 1 to 6 and 14 unchanged; claims 7 to 13 cancelled; new claims 15, 16 and 17 added." or
"Claims 7 to 13 cancelled; new claims 15, 16 and 17 added; all other claims unchanged."

4. [Where various kinds of amendments are made]:
"Claims 1-10 unchanged; claims 11 to 13, 18 and 19 cancelled; claims 14, 15 and 16 replaced by amended claim 14; claim 17 subdivided into amended claims 15, 16 and 17; new claims 20 and 21 added."

"Statement under article 19(1)" (Rule 46.4)

The amendments may be accompanied by a statement explaining the amendments and indicating any impact that such amendments might have on the description and the drawings (which cannot be amended under Article 19(1)).

The statement will be published with the international application and the amended claims.

It must be in the language in which the international application is to be published.

It must be brief, not exceeding 500 words if in English or if translated into English.

It should not be confused with and does not replace the letter indicating the differences between the claims as filed and as amended. It must be filed on a separate sheet and must be identified as such by a heading, preferably by using the words "Statement under Article 19(1)."

It may not contain any disparaging comments on the international search report or the relevance of citations contained in that report. Reference to citations, relevant to a given claim, contained in the international search report may be made only in connection with an amendment of that claim.

Consequence if a demand for international preliminary examination has already been filed

If, at the time of filing any amendments under Article 19, a demand for international preliminary examination has already been submitted, the applicant must preferably, at the same time of filing the amendments with the International Bureau, also file a copy of such amendments with the International Preliminary Examining Authority (see Rule 62.2(a), first sentence).

Consequence with regard to translation of the international application for entry into the national phase

The applicant's attention is drawn to the fact that, where upon entry into the national phase, a translation of the claims as amended under Article 19 may have to be furnished to the designated/elected Offices, instead of, or in addition to, the translation of the claims as filed.

For further details on the requirements of each designated/elected Office, see Volume II of the PCT Applicant's Guide.

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149

PATENT COOPERATION TREATY

PCT

INTERNATI NAL SEARCH REPORT

(PCT Article 18 and Rules 43 and 44)

Applicant's or agent's file reference	FOR FURTHER ACTION	see Notification of Transmittal of International Search Report (Form PCT/ISA/220) as well as, where applicable, item 5 below.
29471/36818		
International application No.	International filing date *(day/month/year)*	(Earliest) Priority Date *(day/month/year)*
PCT/US 01/ 23211	23/07/2001	04/08/2000
Applicant		
ZENITH ELECTRONICS CORPORATION		

This International Search Report has been prepared by this International Searching Authority and is transmitted to the applicant according to Article 18. A copy is being transmitted to the International Bureau.

This International Search Report consists of a total of 3 sheets.

[X] It is also accompanied by a copy of each prior art document cited in this report.

1. **Basis of the report**

   a. With regard to the **language**, the international search was carried out on the basis of the international application in the language in which it was filed, unless otherwise indicated under this item.

   [ ] the international search was carried out on the basis of a translation of the international application furnished to this Authority (Rule 23.1(b)).

   b. With regard to any **nucleotide and/or amino acid sequence** disclosed in the international application, the international search was carried out on the basis of the sequence listing :

   [ ] contained in the international application in written form.

   [ ] filed together with the international application in computer readable form.

   [ ] furnished subsequently to this Authority in written form.

   [ ] furnished subsequently to this Authority in computer readable form.

   [ ] the statement that the subsequently furnished written sequence listing does not go beyond the disclosure in the international application as filed has been furnished.

   [ ] the statement that the information recorded in computer readable form is identical to the written sequence listing has been furnished

2. [ ] **Certain claims were found unsearchable** (See Box I).

3. [ ] **Unity of invention is lacking** (see Box II).

4. With regard to the **title**,

   [X] the text is approved as submitted by the applicant.

   [ ] the text has been established by this Authority to read as follows:

5. With regard to the **abstract**,

   [X] the text is approved as submitted by the applicant.

   [ ] the text has been established, according to Rule 38.2(b), by this Authority as it appears in Box III. The applicant may, within one month from the date of mailing of this international search report, submit comments to this Authority.

6. The figure of the drawings to be published with the abstract is Figure No. 1

   [ ] as suggested by the applicant.

   [ ] None of the figures.

   [ ] because the applicant failed to suggest a figure.

   [X] because this figure better characterizes the invention.

INTERNATIONAL SEARCH REP RT

International Application No
PCT/US 01/23211

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04N7/24

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04N H03D H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)
EPO-Internal, PAJ, WPI Data

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
E	WO 01 78392 A (ADC BROADBAND WIRELESS GROUP I) 18 October 2001 (2001-10-18) abstract page 24, line 15 -page 29, line 4 figure 7 ---	1-5, 15-19, 29-35,40
P,X	WO 00 49710 A (SARNOFF CORP) 24 August 2000 (2000-08-24) abstract page 4, line 1 -page 6, line 15 figures 1,2 ---	1-3,16, 17,30,40
	-/--	

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

* Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
12 March 2002	20/03/2002

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL - 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Hampson, F

2

Form PCT/ISA/210 (second sheet) (July 1992)

150

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 01/23211 151

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	EP 0 853 409 A (HARRIS CORP) 15 July 1998 (1998-07-15)	30,33,40
A	abstract page 2, line 11 -page 2, line 31 page 4, line 15 -page 4, line 34 figures 1,5	1,2,6,7, 16,17, 20,21,29
X	US 4 617 537 A (NOSSEK JOSEF A) 14 October 1986 (1986-10-14)	30
A	column 1, line 59 -column 2, line 4 column 3, line 59 -column 5, line 22 figure 2A	1,16
A	US 5 696 796 A (POKLEMBA JOHN JAMES) 9 December 1997 (1997-12-09) column 5, line 35 -column 6, line 2 figure 4	1,16,30
A	WO 99 40721 A (ZENITH ELECTRONICS CORP) 12 August 1999 (1999-08-12) abstract page 3, line 8 -page 4, line 15 figures 1,2	1,16,30
A	EP 0 323 675 A (PHILIPS ELECTRONICS UK LTD ;PHILIPS NV (NL)) 12 July 1989 (1989-07-12) abstract column 2, line 31 -column 3, line 32	1,16,30

2

INTERNATIONAL SEARCH REP RT

Information on patent family members

International Application No
PCT/US 01/23211

157

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
WO 0178392 A	18-10-2001	AU 4988801 A	23-10-2001
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(74) Agent: LOWE, Allan, M.; Lowe Hauptman Gopstein Gilman & Berner, LLP, Suite 310, 1700 Diagonal Rd., Alexandria, VA 22314 (US).

(81) Designated States *(national)*: AE, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, CA, CH, CN, CR, CU, CZ, DE, DK, DM, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, NO, NZ, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TR, TT, TZ, UA, UG, UZ, VN, YU, ZA, ZW.

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Published:
without international search report and to be republished upon receipt of that report

For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

WO 01/78392 A2

(54) Title: REDUCED BANDWIDTH TRANSMITTER METHOD AND APPARATUS

(57) Abstract: A reduced bandwidth 8VSB, ATSC transmitter transmits to conventional DTV receivers having bandpass filters with root raised cosine (RRC) band edge responses. The transmitter includes a vestigial sideband modulator for transferring energy in a portion of a spectrum on a first side of a predetermined frequency to a second side of the predetermined frequency. At the low band edge of the spectrum, the predetermined frequency is the pilot carrier frequency. At the high end of the spectrum, the predetermined frequency is the Nyquist sampling frequency. The energy transfer is such that no energy is transmitted between at least one band edge of the spectrum and an intermediate frequency between the band edge and the predetermined frequency. The energy is transferred on a frequency to frequency basis as a function of the energy in the RPC transition and the amplitude of a weighting function extending between the intermediate frequency and the predetermined frequency.

REDUCED BANDWIDTH TRANSMITTER METHOD AND APPARATUS

Field of the Invention

The present invention relates generally to transmitter methods and apparatus and, more particularly, to a transmitter method and apparatus wherein energy in a portion of a spectrum on a first side of a predetermined frequency is transferred to a second side of the predetermined frequency in such a manner that at least some of the transferred energy results in a reduction of transmitted bandwidth and receivers responsive to the transmitter respond to the transferred energy as if it were on the first side of the predetermined frequency.

Background Art

The problem of frequency spectrum crowding has become so severe that government allocated spectra for different types of transmission results, in some cases, in overlapping spectra for different transmitters. For example, the United States Federal Communications Commission (FCC) in establishing a table of channel allocations for the Advanced Television Standards Committee (ATSC), eight level vestigial sideband (8VSB) digital television (DTV) transmitters has created severe technical problems for television channel 14 (470–476 MHz). There are terrestrial mobile radio transmitters which the FCC has licensed at 469.975 MHz, i.e., at a frequency allocation very close to the lower band edge of channel 14. It is extremely difficult to protect the low power 469.975 MHz radio emission from the high power DTV signal having energy at 470 MHz, only 0.025 MHz away from the 469.975 MHz emission. This is particularly the case since the FCC has required the 470 MHz band edge to be down only 36 dB from the mid-band amplitude. If prior techniques were used, the channel 14 allocation would be useless in most cases because it would overpower the 469.975 MHz emission, a situation the FCC will not permit.

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Another situation that is problematic is the so-called "N+1" allocation. In the "N+1" allocation, a National Television Systems Committee (NTSC) licensee is assigned a DTV channel that is the next channel up from the licensee's NTSC channel. In this situation, the NTSC aural carrier has significant sidebands that extend to within about 75 kHz of the NTSC channel edge, and therefore to within about 75 kHz of the DTV signal. This creates problems for stations that wish to use frequency selective combiners with separate NTSC and DTV transmitters. Yet another problem is the radio telescope allocation at channel 37 in the UHF DTV band (608–614 MHz). DTV stations on channel 36 or 38 may interfere with radio telescope operations on channel 37.

A typical prior art digital television transmitter adapted to transmit signals containing information indicative of digitally encoded video and aural signals for deriving an ATSC A/53 standard signal is illustrated in Fig. 1 as including multi-bit digital baseband television signal source 10 which drives the cascaded combination of data randomizer 11, Reed-Solomon encoder 12, data interleaver 14, trellis encoder 16 and multiplexer 18. In the 8VSB system, the signals derived from source 10, randomizer 11, encoders 12 and 16, as well as interleaver 14 and multiplexer 18 include three parallel bits having a symbol rate (i.e., sampling frequency) of $\frac{1.539x10^9}{143}$, i.e., the encoded television signal that source 10 derives is sampled 10,762,237.76 times per second. Because each symbol includes two or three bits, the bit rate is substantially higher than the symbol rate. The three parallel bits represent 8 amplitude levels of the encoded television signal.

Multiplexer 18, in addition to being responsive to the output of trellis encoder 18, responds to segment synchronizing source 20 and field synchronizing source 22 to derive an output having the same number of bits as applied to the multiplexer by encoder 16. Multiplexer 14 supplies a multi-bit output signal to pilot inserter 24 which inserts a constant bit pattern representing a DC amplitude equal to 0.625 of the weight of a single bit in the incoming 3 bit pattern. Pilot

inserter 24 derives a multi-bit output signal which drives pre-equalizer filter 26. Pilot inserter 24 derives a multi-bit output signal which drives pre-equalizer filter 26. Pre-equalizer filter 26 supplies a multi-bit intermediate frequency (IF) signal to vestigial sideband modulator or generator 28. Generator 28 feeds a multi-bit digital IF signal to digital to analog converter 30, which supplies an analog IF signal to frequency up converter 32. Converter 32 is a frequency synthesizer that heterodynes the IF output frequency of converter 30 to a radio frequency (RF) carrier frequency. Up converter 32 also inverts the IF spectrum digital to analog converter 30 derives so up converter 32 converts (1) the lowest frequencies in the IF spectrum into the highest frequencies in the RF spectrum and (2) the highest frequencies in the IF spectrum to the lowest frequencies in the RF spectrum. RF up converter 32 applies the modulated carrier frequency signal to antenna 34 via power amplifier 36.

The output signal of digital to analog converter 30 includes orthogonal I and Q channels or components. At predetermined time intervals, after the receiver's root raised cosine filtering, the I channel has one of multiple levels, corresponding to the number of amplitude levels in the 3 bit signal source 10 derives. The Q channel contains no independent information, but causes part of the unwanted lower sideband appearing at the output of up converter 32 to be reduced substantially to zero amplitude. The unwanted lower sideband is removed by circuitry included in vestigial sideband generator 28 and up converter 32 does not reintroduce it. Because up converter 32 "flips" (i.e., inverts) the IF spectrum that digital to analog converter 30 derives, the upper sideband RF output of converter 30 is reduced substantially to zero.

To enable digital to analog converter 30 to produce the desired vestigial sideband signal, vestigial sideband modulator or generator 28 derives the baseband vestigial sideband spectrum illustrated in Fig. 2. The spectrum of Fig. 2 has a 6 MHz bandwidth and includes the 309.44056 kHz pilot carrier frequency that pilot inserter 24 derives, as well as a vestigial sideband that extends 309.4405594 kHz below the pilot carrier frequency, to the left of the carrier as

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illustrated in Fig. 2. The response curve of Fig. 2 has a flat portion 38 that extends throughout the vast majority of the transmitter 6.0 MHz bandwidth. The response curve has monotonic root raised cosine (RRC) transitions 40 and 42 at its upper and lower band edges. Transitions 40 and 42 are symmetrical, each having a frequency extent of about 619 kHz (i.e., twice the pilot carrier frequency) between the edges of the transmitter bandwidth and the corner frequencies where the transitions end and flat portion 38 begins. Receivers responsive to transmitters of the type illustrated in Fig. 1 include a filter with the same response as Fig. 2.

The amplitude response of each of transitions 40 and 42 at any frequency (f) from the mid-point frequency (f_t) of each transition is:

$$Rrc(f) = sqrt\left(\frac{1}{2}\left[1+\cos\pi\left(\frac{f-f_t}{2f_t}\right)\right]\right) \qquad (1)$$

Vestigial sideband modulator 28 for deriving the ATSC A/53 standard has generally used a Hilbert filter or phasing method. My co-pending commonly assigned application, Serial No. 09/239,668, filed January 29, 1999, entitled "Vestigial Sideband Generator Particularly for Digital Television," discloses a modified Weaver modulator as the device for generating the vestigial sideband spectrum that modulator 28 derives.

The ATSC standard 8VSB transmission system, like many other digital transmission systems, includes a certain amount of "excess bandwidth." Because the 8VSB transmitter transmits symbols at a rate of 10.76223776...million per second, the transmitter requires a minimum theoretical bandwidth of 5.381118881...MHz (half the symbol rate). Although this is a theoretical minimum, it is physically impossible to build a transmitter that only uses the minimum bandwidth. As a practical matter, it is necessary to have a certain amount of additional ("excess") bandwidth for filter transition bands. Consequently, the 8VSB system has about 11.5% excess bandwidth. In other words, the 6 MHz channel width divided by 5.381118881...MHz is approximately 11.5% greater than unity. To achieve an overall flat amplitude response between a DTV transmitter and a DTV receiver, the ATSC standard

requires the shape of filtering in transition bands of the edges of the spectrum to have the root raised cosine (RRC) response of Equation (1). Because the RRC filter response is applied twice, at the transmitter and at the receiver, the 8VSB signal magnitude response is squared, providing an overall raised cosine shape. When the receiver demodulates the signal it receives from the transmitter, the I channel of the 8VSB signal has a flat amplitude response.

In the Hilbert filter method, vestigial sideband modulator 28 generates a double sideband signal that is filtered to produce a vestigial sideband signal at an IF of about 10 MHz. Sidebands extend equally around the 10 MHz IF in accordance with:

$$0.5F_{sym} + F_{pilot} = 6MHz - F_{pilot} \qquad (2)$$

where F_{sym} is the 10.76223776...MHz symbol clock frequency of the bits source 10 derives in accordance with the ATSC A/53 standard, and

$$F_{pilot} = \frac{59}{3} FH_{NTSC} = \frac{59}{3} \cdot \frac{4.5MHz}{286}$$

where FH_{NTSC} is the NTSC horizontal line frequency.

Based on the foregoing, a vestigial sideband modulator 28 including a Hilbert transform has a double sideband modulator with sidebands that extend ±5.690559441...MHz on either side of the about 10 MHz carrier. A convenient sampling frequency is four times the 10.76223776...MHz symbol clock rate, i.e., 43.04895105...MHz.

The Hilbert transform modulator uses a phasing method that partially cancels the unwanted sideband of the double sideband signal with the RRC transition. Obtaining a proper root-raised cosine response for vestigial sideband shaping at the 43.04895105...MHz sampling rate of the Hilbert transform requires a finite impulse response (FIR) filter having about 2048 filter coefficients. Implementation of such a filter is difficult.

The Hilbert transform can easily generate a vestigial sideband signal such that DC is 6 dB down with respect to the sidebands. This is because the response of any Hilbert transform approximation is always zero at DC. With only one of

the I and Q modulators included in such a vestigial Hilbert transform sideband generator contributing at DC, the vector sum of the outputs of the two modulators drops in half at DC relative to the vector sum at a frequency where both the I and Q channels contribute to the generator output. However, in the ATSC A/53 standard, the requirement for the root-raised cosine response places the DC output at –3 dB instead of –6 dB. Therefore, the Hilbert transform method of vestigial sideband digital television modulation requires a low frequency equalizer to produce a +3 dB "shelf" at DC and the low frequency portions of the baseband spectrum.

To achieve the ATSC A/53 standard, the Hilbert transform vestigial sideband generator has a linear phase requirement. Consequently, equalizer filter 26 is generally implemented as a finite impulse response filter having a large number of coefficients. Further, a –3 dB requirement exists at the Nyquist frequency of the symbols, i.e., half the symbol frequency, with certain modifications. Hence, equalizer 26 must include a high frequency portion operating at a sampling frequency higher than twice the symbol rate to avoid aliasing, i.e., insertion of information at frequencies that do not exist in the sample frequency due to sampling at a frequency less than twice the highest frequency component being filtered. In this case, the highest frequency being filtered is 5.690559441...MHz, which is more than half the symbol rate. This is another reason why the Hilbert transform method of producing a vestigial sideband signal with root-raised cosine sideband shaping is also quite difficult to implement.

Because power amplifier 34 has a non-linear amplitude response, nonlinear equalizer 26 must apply a substantial non-linear correction to the signal applied to it. Because of the possibility of aliasing and spectral folding through zero frequency, the amount of nonlinear correction which can be applied at 10 MHz is limited, resulting in distortion in the transmitted signal.

My co-pending application is based on my realization that the non-linear correction can be more effectively provided by using the modified Weaver modulator in a way that substantially reduces distortion in the transmitted signal

by employing an IF digital signal having a frequency approximately twice the approximately 10 MHz frequency of the prior art digital IF. The arrangement I previously invented and disclosed in my co-pending application enables the digital IF signal to have a frequency of approximately or exactly 21.5 MHz.

The modified Weaver vestigial sideband modulator includes an oscillator with a folding frequency and lowpass filters with cut off frequencies such that "negative" frequencies the modulating baseband signal pilot inserter 24 derives appear in a signal derived from an adder as a vestige of the opposite sideband, i.e., the positive frequencies of the modulating baseband signal. The folding frequency and the cutoff frequency of the lowpass filters are selected such that a desired amount of the opposite sideband is eliminated but a certain portion of it is passed. Preferably, vestigial sideband modulation is produced by using a Weaver modulator having a reduced folding frequency and lowpass filters having increased cutoff frequencies.

It is, accordingly, an object of the present invention to provide a new and improved method of and apparatus for reducing the bandwidth of a transmitted signal without affecting a conventional receiver responsive to the transmitted signal.

Another object of the invention is to provide a new and improved method of and apparatus for generating a vestigial sideband signal, particularly a digital vestigial sideband IF signal.

An additional object of the invention is to provide a new and improved digital television transmitter apparatus and method wherein a tendency for the transmitted signal in some channels to interfere with signals having a neighboring spectrum is overcome.

A further object of the invention is to provide a new and improved digital television transmitter apparatus and method wherein a tendency for the transmitted signal in some channels to interfere with signals derived by neighboring low power transmitters is avoided.

BNSDOCID <WO 0178392A2 I >

An added object of the invention is to provide a new and improved 8VSB transmitter employing digital filtering techniques for modifying the 8VSB spectrum in such a manner as to avoid interference with neighboring transmitters, wherein the 8VSB transmitter includes passive analog channel output filters and combiners and standard DTV receivers respond to the modified 8VSB spectrum to provide a television signal that is not degraded.

Summary of the Invention

According to one aspect of the invention, a method of preventing substantial interference between portions of first and second spectra that tend to overlap is provided. The overlapping portion of the first spectrum is in a band of frequencies on a first side of a predetermined frequency of the first spectrum. The predetermined frequency is outside the second spectrum and within the first spectrum. The method includes the steps of modifying the first spectrum prior to transmission thereof by transferring energy in frequencies on the first side of the predetermined frequency of the first spectrum to frequencies on the second side of the predetermined frequency so that all energy between a first frequency at a band edge of the first spectrum and a second frequency between the first frequency and the predetermined frequency is reduced substantially to zero. The frequencies on the second side are within the first spectrum. The transfer is such that a receiver designed to respond to the first spectrum causes the transferred energy to be shifted from the second side of the predetermined frequency to the first side of the predetermined frequency. The amount of energy transferred is such that there is no substantial interference between the transmitted first and second spectra.

The receiver includes a bandpass filter with a transition having a predetermined amplitude versus frequency response between a frequency at one band edge of the bandpass filter and a mid-range portion of the bandpass filter response. The predetermined frequency is displaced toward the frequency at said one band edge and from a corner frequency between the mid-range portion and the transition. The energy transfer is preferably such that (1) all energy between

the first and second frequencies is transferred on a frequency to frequency basis to frequencies that are on the second side of the predetermined frequency and are equally displaced from the predetermined frequency, and (2) some energy between the second frequency and the predetermined frequency is transferred, on a frequency to frequency basis, to frequencies that are on the second side of the predetermined frequency and are equally displaced from the predetermined frequency. The amount of the energy transferred from between the second frequency and the predetermined frequency is determined by the relative amplitudes of (a) an amplitude versus frequency weighting function that extends over a band of frequencies between the second frequency and the predetermined frequency and (b) the amplitude versus frequency response of the receiver transition.

The invention is particularly useful for situations wherein the first spectrum is a DTV vestigial sideband spectrum and the receiver bandpass filter has a flat mid-range amplitude versus frequency response and a root raised cosine transition between the corner frequency and the first frequency. In such a case, the predetermined frequency at the low frequency band edge is a baseband pilot carrier frequency and at the high frequency band edge is a Nyquist sampling frequency.

In one embodiment, the modified spectrum has a symmetrical amplitude versus frequency response with respect to its center frequency. In another embodiment, the modified spectrum has an amplitude versus frequency response at its high frequency band edge including the transferred energy and an amplitude versus frequency response at its low frequency band edge that is substantially the same as the receiver transition. In a further embodiment, the modified spectrum has an amplitude versus frequency response at its low frequency band edge including the transferred energy and an amplitude versus frequency response at its high frequency band edge that is substantially the same as the receiver transition.

Preferably, the energy transfer includes lowpass filtering a signal resulting from a baseband signal containing information to be transmitted. The lowpass

BNSDOCID <WO 0178392A2 I >

filtering includes multiplying the signal resulting from the baseband signal by (a) a constant amplitude from DC to another frequency where the transition begins, (b) amplitudes greater than the constant amplitude for frequencies between the another frequency and a further frequency, (c) finite amplitudes less than the constant amplitude for frequencies greater than the further frequency and (d) zero for frequencies that are displaced from the another frequency by the same amount that said frequency at the band edge of the bandpass filter is displaced from the corner frequency.

In one embodiment, the energy transfer includes multiplying a baseband signal containing information to be transmitted by sine and cosine representations to derive a pair of product signals. The sine and cosine representations are at a folding frequency. Replicas of the product signals are lowpass filtered as described *supra*. The signals resulting from the lowpass filtering are combined to derive the reduced bandwidth spectrum.

In one embodiment, the replicas on which the lowpass filtering steps are performed result directly from the multiplying operations to cause the spectrum resulting from the combining step to be symmetrical with respect to a frequency at the center of the first spectrum.

According to another embodiment, second lowpass filtering steps are performed on replicas of the product signals. The second lowpass filtering steps are performed on the product signal replicas before the first mentioned lowpass filtering operation is performed. The second lowpass filtering operations include multiplying the replicas of the product signal by (a) a constant amplitude from DC to the another frequency, and (b) the receiver transition amplitude versus frequency response for frequencies between the another frequency and a frequency removed from the another frequency by an amount equal to the displacement of the band edge of the receiver bandpass filter from the corner frequency. The spectra resulting from the second lowpass filtering steps are shifted to cause the spectrum resulting from the combining step to be asymmetrical with respect to a frequency at the center of the first spectrum. The

asymmetrical spectrum has the same amplitude versus frequency response at a first band edge as the receiver bandpass filter and has an amplitude versus frequency response at a second band edge determined by the first mentioned lowpass filtering operation.

In another embodiment, the signal resulting from the first mentioned lowpass filtering step is multiplied by sine and cosine representations to derive a pair of product signals. The sine and cosine representations are at a folding frequency. Second lowpass filtering operations are performed on replicas of the product signals. The second lowpass filtering operations are performed on the product signal replicas before the first mentioned lowpass filtering operation is performed. The second lowpass filtering operations multiply the replicas of the product signal by (a) a constant amplitude from DC to the another frequency, and (b) the receiver transition amplitude versus frequency response for frequencies between the another frequency and a frequency removed from the another frequency by an amount equal to the displacement of the band edge of the receiver bandpass filter from the corner frequency,

The signals resulting from the second lowpass filtering step are preferably multiplied, according to one embodiment, by cosine and sine representations of an IF to derive second product signals. The second product signals are combined to derive an asymmetrical spectrum that has the amplitude versus frequency response of the receiver transitions at its low frequency edge and the amplitude versus frequency response resulting from the first mentioned lowpass filter operation at its high frequency edge.

According to a further embodiment, I and Q channel Hilbert transform operations are performed on a signal resulting from the baseband signal so that signals resulting from the I and Q channel transform operations have the first mentioned lowpass filtering operations applied to them. Replicas of the signals resulting from the I and Q channel Hilbert transform operations are combined to derive the first spectrum. The combining step preferably includes multiplying the signals resulting from the I and Q channel Hilbert transform operations with

representations of cosine and sine waves at an IF to form product signals that are added.

A further aspect of the invention relates to a transmitter for a first spectrum having a portion that tends to overlap in an interfering manner with a neighboring spectrum. The overlapping portion is in a band of frequencies on a first side of a predetermined frequency that is outside the second spectrum and within the first spectrum. The transmitter includes a filter for modifying the first spectrum prior to transmission. The filter is arranged for transferring energy in frequencies on the first side of the predetermined frequency of the first spectrum to frequencies on a second side of the predetermined frequency. The energy transfer causes substantial elimination from the first spectrum of all energy between a band edge of the first spectrum and another frequency between the band edge and the predetermined frequency. The frequencies on the second side are within the first spectrum. The transfer is such that receivers designed to respond to the first spectrum cause the transferred energy to be shifted from the second side of the predetermined frequency to the first side of the predetermined frequency. An output device is connected to be responsive to the modified spectrum.

The filter preferably includes a digital vestigial sideband modulator connected to be responsive to a digital baseband television signal source and a carrier pilot inserted on the digital baseband television signal source.

In one embodiment, the digital vestigial sideband modulator includes a digital filter including a Hilbert transform combined with a lowpass filter having an amplitude versus frequency response for causing the energy transfer.

In a second embodiment, the vestigial sideband modulator includes a digital folding frequency source for deriving orthogonal components at the predetermined frequency. A digital multiplier arrangement responds to the orthogonal components and the digital signal source for deriving a pair of orthogonal product digital signal components. A digital lowpass filter arrangement responds to the pair of orthogonal product digital signal components.

The orthogonal digital signal components passed by the digital lowpass filter arrangement are combined. The digital lowpass filter arrangement has a response and the digital signal components passed by the digital lowpass filter are combined so the energy transfer occurs and there is no substantial energy between the another frequency and the band edge of the first spectrum.

A further aspect of the invention relates to a memory for controlling an electromagnetic wave transmitter that is arranged for transmitting a signal to a receiver having a predetermined response. The memory stores signals for preventing substantial interference between portions of first and second spectra that tend to overlap. The overlapping portion of the first spectrum is in a band of frequencies on a first side of a predetermined frequency of the first spectrum. The predetermined frequency is outside the second spectrum and within the first spectrum. The stored signals cause a modification of the first spectrum prior to transmission thereof by transferring energy in frequencies on the first side of the predetermined frequency of the first spectrum to frequencies on a second side of the predetermined frequency so that all energy between a first frequency at the band edge and a second frequency between the first frequency and the predetermined frequency is reduced substantially to zero. The frequencies on the second side are within the first spectrum. The transfer is such that a receiver designed to respond to the first spectrum causes the transferred energy to be shifted from the second side of the predetermined frequency to the first side of the predetermined frequency.

The above and still further objects, features and advantages of the present invention will become apparent upon consideration of the following detailed descriptions of several specific embodiments thereof, especially when taken in conjunction with the accompanying drawings.

BNSDOCID <WO 0178392A2 I >

Brief Description of the Drawings

As described, Figs. 1 and 2 are respectively a block diagram of a prior art DTV transmitter and the baseband amplitude versus frequency response curves of the prior art transmitter;

Figs. 3 and 4 respectively include response curves at the lower and upper edges of (1) the root raised cosine response of receivers responsive to a transmitter of the present invention, (2) weighting functions to achieve a reduced bandwidth spectrum, and (3) reduced bandwidth responses resulting from the weighting function;

Fig. 5 includes plots, in dB versus frequency, of the response of receivers and prior art transmitters including root raised cosine transitions and an amplitude versus frequency response of a transmitter including the present invention;

Fig. 6 is a block diagram of a vestigial modulator 28, Fig. 1, including a basic bandpass filter that can theoretically be used to obtain the reduced bandwidth responses of Figs. 3 and 4;

Fig. 7 is a block diagram of a modified Weaver modulator that can actually be employed as vestigial modulator 28, Fig. 1, and which produces the reduced bandwidth responses of Figs. 3 and 4;

Fig. 8 includes an amplitude versus frequency response curve of low-pass filters included in Fig. 6;

Fig. 9 is a block diagram of a modified Weaver modulator that can be employed as vestigial modulator 28 and which has asymmetrical reduced bandwidth edges on only one of the upper or lower frequency edges of the spectrum;

Fig. 10 is a block diagram of a modified Weaver modulator that can be employed as vestigial modulator 28 and which is based on the high frequency transition curve of Fig. 4, but which does not provide reduced bandwidth on the low frequency transitional part of the response of Fig. 2;

Fig. 11 is a block diagram of a Hilbert transform modulator including the reduced bandwidth responses of Figs. 3 and 4;

Fig. 12 includes amplitude versus frequency response plots of the I and Q channels of a transmitter including a vestigial sideband Hilbert transform modulator without the reduced bandwidth features of the present invention;

Fig. 13 is a plot of amplitude versus frequency responses of the I and Q channels of a transmitter including the vestigial sideband Hilbert modulator of Fig. 11, i.e., with the reduced bandwidth features of the present invention; and

Fig. 14 is a block diagram of a computer based system loading a programmable integrated circuit chip to establish digital circuitry for performing operations associated with a digital television transmitter including the present invention.

Detailed Description of Drawings

Before discussing the various embodiments illustrated in the drawings, the voltage amplitude versus frequency response curves of Figs. 3 and 4, respectively at the low and high frequency edges of an 8VSB, ATSC, DTV baseband spectrum, are described to explain the operation of vestigial sideband modulator 28 using principles of a preferred embodiment of the method of the invention.

In Fig. 3, standard RRC transition 42 has (1) a zero amplitude response value at the baseband low frequency edge, where frequency from channel edge is 0 Hz, (2) a monotonically increasing value indicated by Equation (1), and (3) a final amplitude response value of 1.0 at the end of the transition, where frequency from channel edge is 618.8811988 kHz, i.e., twice the pilot frequency of 309.4405994 kHz (indicated by line 44) from the channel edge. In accordance with one aspect of the present invention, the bandwidth of transition 42 is reduced by establishing a guard band between the channel edge and a predetermined frequency removed from the channel edge. All the energy that is within transition 42 and is in the guard band is reduced to zero and transferred to another part of the baseband spectrum, between the pilot frequency and the corner frequency at the end of the transition and the beginning of the flat mid-range portion 38 of Fig. 2. The energy transfer is

performed in such a way as to be recovered by 8VSB, ATSC, DTV receivers as if all of that energy had been in the guard band.

In Fig. 3 the guard band extends from the channel edge to 184,440.5 Hz from the channel edge (i.e., 125 kHz from the 309,440.6 Hz pilot). Hence, all energy in the spectrum that drives modulator 28 between zero and 184,440.5 Hz is transferred, on a frequency to frequency basis, to frequencies above the pilot frequency of 309.4405994 kHz. To make such a transfer, modulator 28 has an amplitude versus frequency response 46 at the low frequency edge of the spectrum. Response 46 has (1) a zero amplitude in the guard band, i.e., all frequencies that are more than 125 kHz lower than the pilot frequency (indicated by line 44), (2) amplitudes lower than RRC transition 42 for frequencies lower than the pilot frequency, (3) the same amplitude as RRC transition 42 at the pilot frequency and (4) amplitudes above RRC transition 42 for frequencies higher than the pilot until transitions 42 and 46 are completed at a low frequency corner frequency that is removed from the channel edge by two times the pilot frequency, i.e., 618.8811988 kHz. For frequencies removed from the channel edge by more than two times the pilot frequency, i.e., for frequencies greater than the low frequency corner frequency, the amplitude of the response is the same as for mid-band portion 38 of Fig. 2 until the high frequency corner frequency is reached; the high frequency corner frequency is the lowest frequency in transition 40 at the high frequency edge of the spectrum. Transition 46 has an amplitude that exceeds the mid-band amplitude for frequencies that are between the corner frequencies and frequencies displaced from the channel edges by about 360 kHz; a peak amplitude of +1 dB is reached at a frequency about 440 kHz from the channel edge.

Response 46 is attained by applying a weighting function (including linear roll off response 48) to RRC response 42 for frequencies in a range between the pilot frequency (indicated by line 44) and frequencies that are 125 kHz lower than the pilot. After the weighting function is applied, additional operations are performed to attain response 46. The weighting function has a zero value for all frequencies between the channel edge and the frequency that is 125 kHz lower than the pilot.

The weighting function of Fig. 3 can be represented by a function G(–f), where –f is the difference between the pilot frequency and every frequency lower than the pilot frequency to the band edge, as

$$G(-f) = 0 \text{ for } f \geq 125 \text{ kHz, and} \qquad (3)$$

$$G(-f) = 1 - \frac{f}{125} \text{ kHz for } 0 \leq f < 125 \text{ kHz.}$$

To arrive at response 46 for every frequency of interest that is less than 125 kHz from the pilot frequency, the difference between response 42 and the weighting function for the frequencies of interest between the channel edge and the pilot is determined. This difference is multiplied by the amplitude of RRC response 42 at the frequencies of interest to obtain a resultant product for each frequency of interest. The resultant product for each frequency of interest is divided by the RRC response at the frequencies that are above the pilot frequency by the same amount that the frequencies of interest are below the pilot frequency, to derive resulting quotient for each frequency of interest. The resulting quotient is added to the RRC response at the frequencies of interest that are above the pilot.

Consider the example of (1) a frequency f_1 that is lower than the pilot by f_2, where f_2 is more than 125 kHz, (2) RRC transition response 42 having a value j at f_1, and (3) RRC transition response 42 having a value of k at the pilot plus f_2. In this example, the product is j^2 and the resultant quotient is $\frac{j^2}{k}$, whereby $\frac{j^2}{k}$ is added to the value k of response 42 at the frequency that is f_2 above the pilot to yield $k + \frac{j^2}{k}$ for the amplitude of response 46 at f_2 above the pilot. Consider a second example of (1) a frequency f_3 that is lower than the pilot by f_4, where f_4 is less than 125 kHz, (2) RRC transition 42 has a value of m at f_3, (3) RRC transition 42 has a value of n at the pilot plus f_4, and (4) the weighting function has a value p at f_3. In

the second example, the product is m (m–p), and the quotient of $\frac{m(m-p)}{n}$ which is added to n to yield $\frac{m(m-p)}{n}$ + n for the amplitude of response 46 at f_4 above the pilot. Illustrative values for responses 42 and 46 for frequencies removed from the pilot by the indicated amounts are given in Table I.

Table I

Frequency removed from the pilot (kHz)	RRC response 42	Response 46
–300	0.023959	0
–200	0.2742153	0
–100	0.5069010	.1013802
– 50	0.6119249	.3671549
– 25	.6608458	.5286766
+ 25	.7505217	.8668987
+ 50	.7909159	.9802923
+ 100	.8620043	1.1004705
+ 200	.9616683	1.0398595
+ 300	.9997129	1.0002871

It is to be understood that other G(f) functions can be used, as long as the G(f) that is employed reduces the bandwidth of the transmitted spectrum sufficiently to prevent interference with transmission from neighboring spectra. The G(f) function must be such that for any voltage at a particular frequency that is removed from the portion of response 42 that is lower than the pilot frequency (i.e., the lower sideband), there is a voltage that is added to a corresponding frequency in the portion of response 42 that is above the pilot frequency. These removed and added voltages are seen by an I channel demodulator of a DTV receiver in the same manner as a transmitted signal conforming with the RRC response curve. Hence, the receiver performance and configuration are unchanged, while achieving a reduction of transmitter bandwidth. In the particular example of Fig. 3, the bandwidth reduction of transition 46 compared to transition 42 is 184 kHz, a sufficient amount to overcome the previously discussed interference problems of the prior art.

The operations described in connection with Fig. 3 can be expressed mathematically as follows:

The RRC response for frequencies less than the pilot (the lower sideband, i.e., LSB) of the receiver to the lower sideband is:

$$Rrc(-f) \quad (4)$$

The reduced bandwidth amplitude of the LSB is:

$$Rrc(-f)*G(-f) \quad (5)$$

The amount of "lost voltage" in the LSB is the difference between expressions (4) and (5), i.e.:

$$Rrc(-f)-Rrc(-f)*G(-f) = Rrc(-f)*[1-G(-f)] \quad (6)$$

Applying the LSB RRC response of the receiver to Equation (6) gives:

$$Rrc(-f)-Rrc^2(-f)*G(-f) = Rrc^2(-f)*[1-G(-f)] \quad (7)$$

Dividing Equation (7) by the RRC response of the receiver for frequencies above the pilot (the upper sideband (USB)) gives:

$$[Rrc(-f)-Rrc^2(-f)*G(-f)]\ Rrc = Rrc^2(-f)*[1-G(-f)]/Rrc(f) \quad (8)$$

The frequency response of the receiver upper sideband becomes:

$$Rrc(-f)+ Rrc^2(-f)*[1G(-f)]/Rrc(-f) \quad (9)$$

Concepts similar to those described for transition 42 at the low frequency edge of the baseband spectrum of Fig. 3 are applicable to transition 40 at the high frequency edge of the spectrum. The ATSC standard 8VSB signal pilot that inserter 24 derives includes essential signal components in the baseband modulating spectrum of Fig. 2 that go all the way to the Nyquist rate. When pilot inserter 24 forms the 8VSB signal, the spectrum (Fig. 2) is flat to 5.071678332...MHz above the 309.4405594 kHz pilot. At that frequency, root raised cosine upper edge band transition 40 begins to roll off the spectrum. At half the symbol rate above the pilot frequency (i.e., 5.381118881...MHz) transition 40 is down 3 dB from the mid-band amplitude indicated by response portion 38. Between this frequency and the channel edge (5.690559441...MHz above pilot) the ideal response transitions from –3 dB to

zero amplitude. Because baseband energy that pilot inserter 24 derives goes all the way to the Nyquist rate, the frequency components in the RRC transition 40 for frequencies just below the Nyquist rate (half the symbol rate or 5.381118881...MHz above pilot, i.e., 5.690559441...MHz) are mirror images of the spectrum for frequencies just above the Nyquist rate. The amplitudes of the mirrored spectra are not the same, but the frequencies are mirrored.

When a receiver responsive to the transmitted 8VSB signal demodulates the signal to symbols sampled at the symbol rate, frequency components within transition region 40 just above the Nyquist frequency alias to frequencies just below the Nyquist frequency. This aliasing at the receiver demodulation is intentional and necessary for the system to work properly.

As was the case in the transition region around the pilot at the low frequency edge of the spectrum, the spectral shape in the transition range at the upper channel edge is modified according to Fig. 4 without affecting the demodulated symbols. Modulator 28 removes energy from the spectrum above the Nyquist rate, and replaces it at the equivalent frequency below the Nyquist rate. This spectral modification considers the shape of the receiver RRC filter.

To make the spectral modification to the upper band edge, the 8VSB Nyquist rate (half the symbol rate plus the pilot frequency) of Fig. 4 replaces the pilot frequency of Fig. 3. References to the lower sideband in the discussion of Fig. 3 are interchanged with references to the upper sideband in the discussion of Fig. 4. In other words, in Fig. 4 energy from the frequency range between the Nyquist rate (5.381118881...MHz plus pilot (5.6904559441 MHz)) and the upper channel edge is removed. Modulator 28 compensates for the loss of energy by increasing the signal amplitude between pilot plus 5.071678322...MHz and pilot plus 5.381118881...MHz.

In the amplitude versus frequency response curves at the upper edge of the spectrum as illustrated in Fig. 4, RRC transition 40 is replaced by a reduced bandwidth transition 50 that is derived with the help of a weighting function including linear rolloff 52 from the Nyquist frequency of 5.690559441 MHz (indicated by line 54) to a frequency 125 kHz above the Nyquist frequency. The weighting function has a zero amplitude from the frequency 125 kHz above the Nyquist frequency to the upper edge of the spectrum, at 5.690559441 MHz above the 309.44056 kHz pilot. Hence, modulator 28 removes voltage from frequencies above the Nyquist frequency, and replaces it at the same frequency below the Nyquist frequency, such that the frequency-transposed energy is the same at the receiver demodulator responsive to the receiver RRC filter. Such action does not change the receiver demodulated I channel signal after it is sampled. This is because the signal components at a frequency, f, displaced above the Nyquist frequency produce the same result as sampling a frequency component, f, that is below the Nyquist frequency.

Fig. 4, essentially a mirror image of the lower sideband response of Fig. 3, shows the detail of how modulator 38 forms the narrowband 8VSB spectrum at the upper channel edge. The weighting function has a value of unity for frequencies below the Nyquist frequency plus the pilot frequency, indicated by line 54. For frequencies inserter 24 derives that are between the Nyquist frequency plus the pilot frequency and 125 kHz above the sum of the pilot and Nyquist frequencies, the weighting function straight line 52 goes from 1.0 to zero. In the remaining range, from 125 kHz above the sum of the pilot and Nyquist frequencies to the upper channel edge, the weighting function has a value of zero. The removed lower sideband energy is multiplied by the RRC response. Then it is frequency inverted above the sum of the pilot and Nyquist frequencies, and divided by the RRC response, and finally added to the RRC response. The result is the reduced band response 50 shown in Fig. 4. The shaping of Fig. 4 results in 184 kHz being removed from the upper sideband. The demodulated I channel of this signal that the

BNSDOCID <WO. 0178392A2 I >

receiver derives is identical to the demodulated I channel signal that the receiver derives in response to the standard RRC 8VSB spectrum of Fig. 2.

While the foregoing descriptions of Figs. 3 and 4 describe a method of producing filter shapes which a DTV receiver having a standard 11.5% excess bandwidth that demodulates correctly, there is no requirement to use any particular filter design method. There are many well-known filter design algorithms which can be used to produce filters to perform the method. A suitable design method is to use the Parks-McClellan filter design procedure, which, in turn, uses the Remez algorithm.

These methods can be applied to only the lower band edge, or only the upper band edge, or to both band edges simultaneously, depending on the purpose for reducing the bandwidth of the 8VSB or other appropriate signal. For example, if the problem is producing an (N+1) combiner system, it may only be necessary to reduce the bandwidth at the lower band edge. Or, if an FM station at 88.1 MHz suffers interference from a new DTV allocation at channel 6, only the upper band edge of the 8VSB signal need be reduced. With the previously described particular G(f) function, there is enough energy removed from the channel edges to make the in-band response at the peaks of responses 46 and 50 in Figs. 3 and 4 rise by approximately 1 dB.

To provide maximum bandwidth reduction on just one channel edge, the bandwidths on both sides of the spectrum can be reduced. In such a case, the resulting spectrum is frequency shifted to one side or the other by the amount of the bandwidth reduction on one side. In these examples, with 184 kHz removed from each side, and with the resulting spectrum shifted by 184 kHz, a total guard band of 368 kHz is created on one side of the spectrum.

Fig. 5 includes amplitude (in dB) versus frequency responses of the prior art spectrum 58 and the spectrum 60 resulting from the present invention. The prior art spectrum 58 includes RRC, monotonic transitions 40 and 42 at the upper and lower edges of the bandpass region. Transitions 40 and 42 conform with the RRC requirement and have amplitudes of –35 dB at the edges of the spectrum and

ENSDOCID <WO 0178392A2 I >

a 0 dB value at the corner frequencies that begin at the mid-range of spectrum 58; the corner frequencies are approximately 600 kHz from each edge of spectrum 58. Spectrum 58 has a 0 dB value throughout its mid-range.

Response 60 includes transitions 62 and 64 at the lower and upper edges of its spectrum. At –35 dB, each of transitions 62 and 64 is displaced toward the mid-portion of response 60 by approximately 184 kHz from the frequencies of transitions 42 and 40. Each of transitions 62 and 64 includes a peak 66 at about +1 dB, i.e., 1 dB above the 0 dB the mid-range amplitude 38 of responses 58 and 60. After having passed through peak points 66, each of transitions 62 and 64 returns to the 0 dB level at a corner frequency where the 0 dB mid-range amplitude of response 60 begins. As described previously in connection with Figs. 3 and 4, the energy in transitions 62 and 64 at those portions of transitions 62 and 64 above transitions 40 and 42 is due to the energy removed from spectrum 60 that is below the energy in transitions 40 and 42.

From a theoretical conceptual standpoint, a bandpass filter having the response curve 60 of Fig. 5 is the easiest way to attain the spectrum of desired reduced bandwidth response 60. Fig. 6 is a block diagram of a vestigial modulator including bandpass filter 74 having amplitude versus frequency response curve 60.

The vestigial modulator illustrated in Fig. 6 includes simple digital multiplier 70 having a first input responsive to the digital sequence that filter 26 derives and a second input responsive to a digital signal representing a 45 MHz cosine wave that IF carrier frequency generator 72 derives. Simple multiplier 70 responds to the pair of bits supplied to it so that for each +1 symbol at the second input, the multiplier output is the same as the symbol at the first input; for each –1 symbol at the second input the multiplier output is inverted in polarity from the first input; for each zero symbol at the second input the multiplier output is zero. Generator 72 generates each cycle of the cosine sequence supplied to multiplier 140 as a digital, tri-level sequence 1, 0, –1, 0, or a bi-level digital sequence 1, 1, 1, –1. Thus generator 72 produces four binary symbols for each cycle of the IF carrier frequency, which has a frequency of 45 MHz. Thus, IF carrier frequency generator

BNSDOCID <WO 0178392A2 I >

72 derives bi- or tri-level symbols at a 180 MHz symbol rate. Multiplier 70 is effectively a mixer that derives equal upper and lower sidebands centered on the IF cosine wave that generator 72 derives. Filter 74 shapes the double sideband output of multiplier 70 to provide vestigial sideband spectrum 60. Spectrum 60, at the output of modulator 28, is up converted and inverted by up converter 32, after being converted into an analog signal by analog to digital converter 30.

While the bandpass arrangement of Fig. 6 is conceptually the simplest, it has numerous disadvantages. Vestigial bandpass filter 74 requires a higher symbol calculation rate than the prior art Hilbert or Weaver vestigial sideband modulators. Bandpass filter 74 requires more filter taps than the Hilbert and Weaver approaches, as well as a higher sampling rate than the Weaver and Hilbert approaches which operate as baseband filters, i.e., between 0 and 6 MHz. Hence, from a practical standpoint, the bandpass filter arrangement of filter 76 would not usually be adopted in an actual digital television transmitter.

Reference is now made to Fig. 7 of the drawing, a block diagram of a modified Weaver modulator for deriving a vestigial sideband digital IF signal having reduced bandwidth response 60, Fig. 5. The structure illustrated in Fig. 7 is one preferred embodiment of vestigial modulator 28, Fig. 1.

A multi-bit parallel digital input signal having 12 bits and a sampling frequency of 10.76223776...MHz, as derived from pilot inserter 24 and pre-equalizer filter 26, in the form of an ATSC/A53 television standard, is applied in parallel to identical simple digital multipliers 140 and 142, of the type described *supra* in connection with multiplier 72, Fig.6. Multipliers 140 and 142 are responsive to orthogonal digital representations of cosine and sine waves each having a 10.76223776...MHz bit frequency, as derived from folding frequency generator 144, which derives sequences similar to the sequence that generator 72, Fig. 5, derives. Thus, the frequency of each cycle of the cosine sequence supplied by generator 144 to multiplier 140 is 2.690559441...MHz. The sine sequence that generator 144 supplies to multiplier 142 is displaced by one bit, i.e., 90°, from the cosine sequence. A bi-level cosine sequence would be $1, 1, -1, -1$, with the

corresponding sine sequence being –1, 1, 1, –1. A tri-level cosine sequence would be 1, 0, –1, 0 with the corresponding tri-level sine sequence being 0, 1, 0, –1.

For clarity to describe and illustrate the invention, the levels derived by generator 144 are shown as being simultaneously supplied to multipliers 140 and 142 in the illustrated embodiment. It is to be understood, however, that the cosine and sine sequences can be interleaved, in which case only a single simple multiplier is necessary; in such a case the inputs and outputs of multipliers 140 and 142 are delayed from each other by one-half the period of the 10.76223776...MHz symbol rate.

The product signal multipliers 140 and 142 derive are respectively applied to identical finite impulse response (FIR) digital lowpass filters 146 and 148. Each of filters 146 and 148 can be implemented with approximately 256 finite impulse response coefficients, selected to produce the reduced bandwidth response 60, including transitions 62 and 64. Hence, filters 146 and 148 pass, without substantial attenuation, the frequency of the sine and cosine representing waves derived by generator 144 and block virtually all components at the high frequency end of the digital television signal supplied to multipliers 140 and 142. The response curves for filters 146 and 148 are described *infra* in connection with Fig. 8. Filters 146 and 148 can be replaced by a single filter that is multiplexed to respond to the two product signals multipliers 140 and 142 derive.

Frequency offset circuitry 169, described in detail *infra*, changes the 21.52447552 MHz output of filters 146 and 148, to a precise round, i.e., decimal, frequency, preferably 21.5 MHz. Frequency offset circuitry 169 supplies to interpolator sets 150 and 152 I and Q channel lowpass filtered sequences that are frequency offset replicas of the sequences digital filters 146 and 148 derive. Interpolator sets 150 and 152 multiply by eight the sampling rate of filters 146 and 148. To this end, interpolator set 150 includes three cascaded times two up sampling interpolators 154–156, and interpolator set 152 includes three cascaded times two up sampling interpolators 158–160. Consequently, the sampling frequency of the output signals of each of interpolators 156 and 160 is 86.0 MHz.

BNSDOCID <WO 0178392A2 I >

The output signals of interpolators 156 and 160 are respectively supplied to identical digital multipliers 162 and 164, preferably constructed the same as multipliers 70, 140 and 142. Multipliers 162 and 164 are respectively responsive to cosine and sine sequences digital IF generator 166 derives. Each cycle that generator 166 derives has the same values as the cosine and sine sequences derived from folding frequency generator 144; each cycle of the sine and cosine sequences supplied to multipliers 162 and 164 has an IF of 21.5 MHz. Because there are four symbol values in each cycle of the cosine and sine representing sequences that generator 166 derives, the 86.0 MHz symbol frequencies that generator 166 derives are the same as the symbol frequencies that interpolators 156 and 160 derive.

The digital output signals of multipliers 162 and 164 are linearly combined, i.e., summed, in digital adder 168 to produce a digital vestigial sideband IF signal having amplitude versus frequency response 60, Fig. 5; the output of adder is centered at 21.5 MHz.

The output signal of adder 168 is applied to digital to analog converter 30 (Fig. 1), which derives an analog IF vestigial sideband signal including I and Q channels such that the I channel, at predetermined times, contains one of eight levels. The eight levels are commensurate with the binary value of the signal multiplexer 18 derives at a time corresponding to each sampling time. The Q channel of digital to analog converter 30 causes part of the unwanted vestigial sideband that modulator 28 derives to be reduced to a zero amplitude.

Frequency offset circuitry 169 includes digital frequency synthesizer 170 (which functions as an offset frequency generator), general purpose digital mutipliers 172, 174, 176 and 178, and digital adders (i.e., linear combiners) 180 and 182 for respectively adding and subtracting the input signals supplied to them. General purpose multipliers actually multiply the magnitudes of the binary signals supplied to them. Adders 180 and 182 respectively drive interpolators 154 and 158. Multipliers 172 and 178 are driven in parallel by the output signal of lowpass filter 146 while multipliers 174 and 176 are driven in parallel by the output signal of lowpass filter 148. Offset frequency generator 170 derives orthogonal cosine and

sine sequences, each cycle of which has digital values that are the same as each cycle of the sequences that folding frequency generator 144 derives.

The cosine output of offset frequency generator 170 is applied in parallel to multipliers 172 and 176, while the sine output of the offset frequency generator is applied in parallel to multipliers 174 and 178. To enable the signal that adder 168 derives to be at an integer value, such as 21.5 MHz, the frequency of generator 170 is set to a value equal to the difference between the desired offset and one-fourth of the bit frequency of the IF output signal; in the specific example, the frequency of offset generator 170 is 24.4755244...kHz.

The digital vestigial sideband modulator of Fig. 7 functions such that the digital output signals of lowpass filters 146 and 148 are respectively represented by $0.5\sin(\omega_m - \omega_4)t$ and $0.5\cos(\omega_m - \omega_4)t$. The signals multipliers 172, 174, 176 and 178 derive are respectively represented by:

$$0.5\cos(\omega_m - \omega_4)\, t^*\cos\omega_o t$$

$$0.5\sin(\omega_m - \omega_4)\, t^*\sin\omega_o t$$

$$0.5\sin(\omega_m - \omega_4)\, t^*\cos\omega_o$$

$$0.5\cos(\omega_m - \omega_4)\, t^*\sin\omega_o t$$

where ω_o is the frequency of the television signal derived by pilot inserter 24, ω_4 is the frequency of generator 144, and ω_o is the frequency offset generator 170 derives.

Adder 180 sums the output signals of multipliers 172 and 174 to derive a signal which can be shown, by trigonometric manipulation, to equal

$$0.5\sin(\omega_m - \omega_{4+}\omega_o)t.$$

Thus, the output signal of adder 180 is frequency shifted upwardly from the output of lowpass filter 146 by ω_o, the offset frequency of generator 170. Adder 182

ENSDOCID <WO 0178392A2 I >

subtracts the output signal of multiplier 178 from the output signal of multiplier 176 to derive a signal which can be shown by trigonometric manipulation, to equal

$$0.5\cos(\omega_m - \omega_{4+}\omega_0)t.$$

Thus, the output signal resulting from the subtracting operation of adder 182 is offset in frequency upwardly from the output signal of lowpass filter 148 by ω_0. The output signals of linear combiners 180 and 182 thus have the same frequency but are phase displaced from each other by 90°. Proper manipulation of the polarities at the inputs of adders 180 and 182 or at frequency generator 170 can reverse the direction of the frequency offset.

It is desirable for the output frequencies of linear combiners 180 and 182 to be a precise numerical value to simplify the frequency synthesizer included in up converter 32 which shifts the IF digital to analog converter 30 derives to the carrier frequency power amplifier 36 derives, without affecting the modulation on the carrier digital to analog converter 30 derives.

Up converter 32 is simplified because digital television signals derived in accordance with the ATSC standard can be offset from the carrier frequency by several predetermined frequencies for various reasons, such as the possibility of interference between signals emitted from antennae that might be spaced from each other by 200 or more kilometers but which are at nominally the same frequency, i.e., have the same television channel number. These "channel offsets" can be programmed into the offsetter of Fig. 7.

Because multipliers 172–178 respond to the output signals of lowpass filters 146 and 148 which are at the symbol frequency of 10.76223776 MHz, multipliers 172–178, which perform simple functions such as passing, blocking or inverting the values supplied to them, can be replaced by a single multiplier that is time multiplexed.

The modulator of Fig. 7 enables the digital IF center frequency to be offset so it has a round number without using high speed general purpose digital multipliers at the high frequency sampling rate of eight times the symbol frequency. Digital multiplication is performed at the high frequency by using simple digital

multipliers 162 and 164. While multipliers 172–178 are of the general purpose type, these multipliers operate at relatively low frequencies in response to the low sampling frequencies derived by filters 146 and 148 and the low frequency outputs of synthesizer 170.

Fig. 8 is a plot of the identical amplitude (in dB) versus frequency responses 181 of FIR lowpass filters 146 and 148, which enable the vestigial sideband modulator of Fig. 7 to provide the amplitude versus frequency response 60, Fig. 5. Response curve 181 is flat, at 0 dB, from DC to a frequency, f_6, equal to the greater of $|f_4 - f_f|$ and $|f_f + f_5|$, where f_4 is the upper edge of the baseband television signal, i.e., 6 MHz, f_f is the folding frequency of generator 144, i.e., one-fourth of the symbol rate of generator 144, and f_5 is the lowest frequency of the baseband spectrum, which for a television spectrum is 0. For frequencies greater than f_6, response 181 has the same shape as response 60, Fig. 5, for all frequencies higher than the mid-range portion of response 60. Thus, response 181 of each of filters 146 and 148 for all frequencies in excess of f_6 is determined in the same way as response 50, Fig. 4.

In a preferred embodiment, lowpass filters 146 and 148 provide responses 181 by employing 256 coefficients. The first 128 of these coefficients are listed in Table II. The remaining 128 coefficients are symmetrical (i.e., the mirror images) with the coefficients of Table II and thus are not repeated.

BNSDOCID <WO 0178392A2 I >

Table II

3.628325E-04	-6.000522E-04	-2.366306E-03	-3.073699E-03
-1.451344E-03	6.857925E-04	8.411454E-04	-4.176315E-04
-6.61839E-04	2.99519E-04	5.384736E-04	-2.617595E-04
-4.5618E-04	2.546355E-04	3.961486E-04	-2.611832E-04
-3.480937E-04	2.749343E-04	3.063179E-04	-2.924158E-04
-2.672116E-04	3.112784E-04	2.280809E-04	-3.308854E-04
-1.876924E-04	3.506789E-04	1.449127E-04	-3.702461E-04
-9.902383E-05	3.891541E-04	4.980066E-05	-4.068929E-04
3.051773E-06	4.234508E-04	-5.91877E-05	-4.384037E-04
1.182099E-04	4.513138E-04	-1.794866E-04	-4.616347E-04
2.42067E-04	4.681326E-04	-3.056173E-04	-4.703579E-04
3.685901E-04	4.667144E-04	-4.301627E-04	-4.553552E-04
4.911573E-04	4.372383E-04	-5.488818E-04	-4.101423E-04
6.02634E-04	3.727492E-04	-6.516444E-04	-3.242859E-04
6.941199E-04	2.621333E-04	-7.319134E-04	-1.899532E-04
7.586739E-04	1.03193E-04	-7.745926E-04	-1.535478E-06
7.778655E-04	-1.149862E-04	-7.645711E-04	2.485024E-04
7.319169E-04	-4.00594E-04	-6.762365E-04	5.737777E-04
5.93837E-04	-7.712424E-04	-4.810862E-04	9.967176E-04
3.341716E-04	-1.25521E-03	-1.501516E-04	1.552136E-03
-7.353534E-05	-1.893648E-03	3.386813E-04	2.286433E-03
-6.461303E-04	-2.737301E-03	9.962169E-04	3.25365E-03
-1.387882E-03	-3.842199E-03	1.819697E-03	4.509306E-03
-2.290122E-03	-5.261099E-03	2.797515E-03	6.102438E-03
-3.342954E-03	-7.040816E-03	3.92574E-03	8.08091E-03
-4.547874E-03	-9.220721E-03	5.234013E-03	1.048834E-02
-5.982622E-03	-1.188374E-02	6.829642E-03	1.343403E-02
-7.816085E-03	-1.517673E-02	9.007661E-03	1.717647E-02
-1.050352E-02	-1.954075E-02	1.246001E-02	2.245214E-02
-1.514291E-02	-2.624444E-02	1.903638E-02	3.157824E-02
-2.512952E-02	-3.993826E-02	3.583379E-02	5.546171E-02
-5.903999E-02	-9.553206E-02	.1446489	.4556426

In certain situations, discussed previously, it is desirable for the reduced bandwidth transmitter of the present invention to include a vestigial sideband modulator 28 having an asymmetrical response. In one embodiment, the asymmetrical response is derived by using the Weaver technique to derive a spectrum having the normal RRC transitions at the upper and lower band edges. The spectrum is then shifted and the technique of the invention is applied to the resulting shifted spectrum to the band edge having the narrower frequency band. For example, if it is desired to reduce the bandwidth of the spectrum only at the high frequency edge of the spectrum, the prior art spectrum with the RRC transitions is shifted to a lower center frequency. The principles of the invention are then applied to the shifted spectrum. The resulting spectrum has an RRC roll-off at its lower edge and a narrow band response, per Fig. 5, at its upper band edge. The asymmetry occurs even though both FIR lowpass filters responsive to the shifted spectrum have the same response per Fig. 8, because of the "negative" frequencies of the Weaver vestigial sideband operations.

Fig. 9 is a block diagram of one arrangement for achieving such an asymmetrical spectrum. Fig. 9 is similar to Fig. 7, except that Fig. 9 includes two additional FIR lowpass filters 183 and 184, respectively responsive to summers 180 and 182. The output signals of lowpass filters 183 and 184 are respectively supplied to interpolators 154 and 158. Another difference between Figs. 7 and 9 is that in the asymmetrical vestigial sideband modulator of Fig. 9, each of lowpass filters 146 and 148 has, at its upper edge, a standard wideband response with RRC transitions beginning at f_6, as indicated by monotonic RRC transition 40, Fig. 5, and a flat 0 dB response from DC to the beginning of the monotonic RRC transition. The frequency of offset oscillator 170 is set to provide the desired offset to assist in achieving the asymmetrical frequency response and is not necessarily selected to achieve a precise IF having a rounded numerical value. For asymmetric spectra having reduced energy at the upper channel edge, the frequency and phase relationships of oscillator 170 are set so that frequencies which eventually translate to the upper channel edge are increased to frequencies

that will pass through filters 183 and 184, and the frequencies which eventually translate to the lower channel edge are decreased to frequencies that will pass through filters 183 and 184. For asymmetric spectra having reduced energy at the lower channel edge, the direction (sign) of the frequency offset is reversed

The frequency responses of lowpass filters 183 and 184 are identical to shape the edge of the spectrum having a reduced bandwidth. To these ends, identical FIR lowpass filters 183 and 184 have the same amplitude versus frequency response curve as illustrated in Fig. 8.

Fig. 10 is a block diagram of a variation of vestigial sideband generator 28, wherein a reduced bandwidth spectrum is provided only at the upper band edge of the spectrum. In the vestigial sideband modulator of Fig. 10, interpolator 190 is connected to respond to the output signal of a pre-equalizer filter 26 to double the sampling frequency of the output of filter 26, i.e., interpolator 190 doubles the symbol rate of the sequence filter 26 derives. FIR lowpass filter 192 responds to the output signal of interpolator 190. Filter 192 has an amplitude versus frequency response, which when multiplied by the standard RRC bandpass response of a DTV receiver results in the upper sideband spectral shape that enables the transmitted spectrum to conform with upper sideband transition 62, Fig. 5. To this end, FIR lowpass filter 192 has the amplitude versus frequency response 181, of Fig. 8.

The output signal of lowpass filter 192 drives multipliers 140 and 142 in parallel. Multipliers 140 and 142 respond to orthogonal sequences generator 144 derives. The resulting product signals at the outputs of multipliers 140 and 142 respectively drive FIR lowpass filters 146 and 148, having standard RRC transitions for all frequencies above f_6. Digital folding frequency generator 144 supplies cosine and sine digital sequences having a symbol rate twice the symbol rate that filter 26 derives. Hence, in the embodiment of Fig. 10, FIR lowpass filter 192 derives the reduced bandwidth response 62, while FIR lowpass filters 146 and 148 provide the RRC monotonic response 42, at the low frequency spectrum edge.

Lowpass filters 146 and 148 respectively drive simple multipliers 162 and 164, without the intermediary of interpolator sets 150 and 152 included in the embodiments of Figs. 7 and 9. Multipliers 162 and 164 also respectively respond to cosine and sine output sequences of digital IF frequency generator 166; generator 166 derives four cosine and four sine symbols per cycle. Multipliers 162 and 164 respond to the inputs thereof to derive product sequences which are added in summer 168, which drives digital to analog converter 30.

The reduced bandwidth concept of the present invention is also applicable to Hilbert transform devices used as vestigial sideband modulator 28. Fig. 11 is a conceptual block diagram of a Hilbert transform type vestigial sideband modulator employed in the prior art and which can be modified in accordance with the present invention by changing the responses of FIR lowpass filters. The Hilbert transform modulator of Fig. 11 includes interpolator 200 which responds to the output of pilot inserter 24 to increase the sampling frequency of the samples pilot inserter 24 derives by a factor of 2, so that interpolator 200 produces samples at a frequency of 21.52447552...MHz. Increasing the sampling frequency of the output of pilot inserter 24 is necessary because the highest frequency to be transmitted by vestigial sideband modulator 28 is 5.690559441...MHz, i.e., the 6 MHz channel width minus the 309.44006 kHz pilot frequency. The 5.690559441...MHz frequency is slightly higher than the Nyquist frequency for symbol rate processing of 10.76223776...MHz.

The output signal of interpolator 200 drives I channel processor 202 and Q channel processor 204 in parallel. I channel processor 202 includes an I channel filter, pre-equalizer filter 26 and a delay element that compensates for the delay introduced by Q channel processor 204. Q channel processor 204 includes a Q channel filter, pre-equalizer filter 26 and a Hilbert transform which introduces a 90° phase shift for all frequencies interpolator 200 supplies to Q channel processor 204. The amplitude of the I channel filter of I channel processor 200 at any frequency the vestigial modulator derives is proportional to the sum of the voltages the transmitter derives in the upper and lower sidebands at that

ENSDOCID <WO _ 0178392A2 I >

frequency. The upper and lower sidebands respectively include those frequencies which are greater and less than the pilot frequency. Similarly, the amplitude response of the Q channel filter of Q channel processor 204 at any frequency the vestigial sideband modulator derives is proportional to the upper sideband voltage minus the lower sideband voltage at that frequency.

Interpolator 200 produces a certain amount of aliasing because the spectrum that pilot inserter 24 derives extends from DC to the Nyquist sampling rate with no guard band beyond the Nyquist rate. The aliasing that interpolator 200 introduces is desirable and must appear as a partial inverted replicated spectrum in the spectral tail the transmitter derives at the upper sideband edge. The responses of interpolator 200 and the I channel filter of I channel processor 200 and the Q channel filter of Q channel processor 204 together determine the spectral shape of the spectral tail at the upper channel edge.

The I channel filter of I channel processor 202 has a linear phase which causes a constant group delay of the signal I channel processor 202 processes, such that the rate of change of phase with respect to frequency is constant. Such a result can be achieved by using a linear phase FIR filter. As indicated by the amplitude (in dB) versus frequency response 206, Fig. 12, the I channel filter of processor 202 includes a +3 dB bump at the low frequency end of the spectrum. The +3 dB bump continues to DC. The +3 dB bump indicated by response 206 represents the difference between the raised cosine response and the root raised cosine response. In the raised cosine response, the pilot frequency is at a frequency that is 6 dB below the 0 dB mid-band response. In the root raised cosine response, the pilot is at a frequency that is down 3 dB from the 0 dB mid-range value. The Q channel filter of Q channel processor 204 has a roll-off that determines the sideband asymmetry, i.e., the level of the vestigial sideband, near the frequency of the pilot carrier.

In the conventional Hilbert transform modulator, the output signals of I and Q channels processors 202 and 204 are respectively supplied to FIR, RRC lowpass filters 208 and 210. Filters 208 and 210 have flat 0 dB responses from

DC to the beginning of the monotonic RRC roll-off, i.e., transition. Beyond the roll-off frequency, filters 208 and 210 have the RRC transition.

Interpolator 200, I channel processor 202, Q channel processor 204, and RRC lowpass filters 208 and 210 are, in actuality, consolidated into only two filters. For the I channel, interpolator 200, lowpass filter 208, and I channel processor 202 are combined in a single polyphase filter. The RRC lowpass filter 208, RRC lowpass filter 210, and interpolator 200 permit a certain amount of the inverted replicated spectrum above half the symbol rate to pass. The channel polyphase filter including interpolator 200, processor 202 and lowpass filter 208 includes two sets of interleaved coefficients which alternate two times during each symbol clock pulse of the output of pilot inserter 24. Similarly, interpolator 200, Q channel processor 204 and lowpass filter 210 are, in actuality, combined into a single polyphase filter.

The output signals of lowpass filters 208 and 210 drive one input of simple multipliers 212 and 214, respectively. The remaining input of multiplier 212 responds to a digital cosine sequence that IF carrier frequency generator 216 derives, while the other input of multiplier 214 responds to a sine wave sequence generator 216 derives. Multipliers 212 and 214 derive product signals which are combined in summer 218.

The output signal of summer 218 has the I and Q channel responses 206 and 220 (Fig. 12), respectively. Q channel response 220 includes RRC roll-off transition 222 at the low frequency edge of the spectrum, such that transition 222 has a value of –35 dB at 559 Hz. Response 220 also includes a flat mid-range 0 dB response 224 and asymmetrical roll-off 226 at the upper edge of the spectrum. The I channel response, in addition to including bump 206, has the same mid-channel response 224 and high edge roll-off 226 as the Q channel.

In the low frequency region of response curve 220 (Fig. 12) corresponding to the vestigial sideband portion of the spectrum, the I channel response 206 and the Q channel transition 222 are unequal. As the Q channel amplitude approaches DC, the resulting spectrum that summer 218 derives approaches a double

sideband. At the frequencies where the I and Q channel responses are the same, i.e., at all frequencies to the right of the intersection of response 206 and transition 222, the spectrum that summer 218 derives has a single sideband.

To enable the Hilbert transform vestigial modulator to produce a reduced bandwidth spectrum in accordance with the present invention, the I and Q channel frequency responses of vestigial sideband modulator illustrated in Fig. 11 are both modified. For a particular reduced bandwidth spectral shape, the I channel Hilbert response is the sum of the upper and lower sideband shapes, while the Q channel response is the difference between the upper and lower sideband shapes or responses. At low frequencies, near the pilot frequency, the I channel response of the Hilbert vestigial modulator is the same for the present invention as for the RRC prior art modulators. The I channel response of the Hilbert modulator in accordance with the invention is different from the RRC transition only if the upper spectrum edge has a reduced bandwidth. Such a result is achieved by changing FIR lowpass filter 208 from an RRC response to the response of Fig. 8. There is no change to the I channel filter in processor 202 so that the I channel response at low frequencies, in accordance with the present invention, is exactly the same as it is in the prior art RRC response.

The Q channel filter in processor 204 has the same response at its low frequency edge as bandpass filter 74 of Fig. 6, as indicated by transition 64 with peak 66 in Fig. 5, as well as by response 46, Fig. 3. Q channel FIR lowpass filter 210 has frequency response 181, Fig. 8, to achieve the bandwidth reduction at the upper edge of the spectrum.

Fig. 13 includes I channel response 220 and the Q channel response 222 of the Hilbert transform modulator of Fig. 11, with modified FIR lowpass filters to achieve the bandwidth reduction of the present invention. The Q channel response 220 includes a 0 dB mid-range portion 224, upper band edge transition 226 and lower band edge transition 228. Transitions 226 and 228 assist in providing the reduced bandwidth and have the same shape as transitions 62 and 64, Fig. 5. Hence, each of transitions 226 and 228 has a peak at 1 dB above 0 dB

mid-range response 224. Transition 228 has an amplitude of –35 dB at 559 Hz, while transition 226 has an amplitude at –35 dB at a frequency of 5.69 MHz. The I channel response is the same as the Q channel response for frequencies greater than the frequency where I channel response 220 intersects the peak of transition 228 at +1 dB. Summer 218 combines the I and Q channel responses of Fig. 13. Summer 218 drives digital to analog converter 30 to provide substantially the same response as response 60, Fig. 5.

The functions of the digital circuitry of Fig. 1, including data randomizer 11, Reed Solomon encoder 12, data interleaver 14, trellis encoder 16, multiplexer 18, segment sync source 20, field sync source 22, pilot inserter 24, pre-equalizer filter 26 and vestigial sideband modulator 28, are stored in flash random access memory 230, Fig. 14. Flash random access memory 230 includes a file that stores the coefficients for the filters, including the coefficients of Table II for the FIR lowpass filters having response 181, Fig. 8. The signals that flash random access memory 230 stores are supplied to computer 234 which controls connections of gates in several field programmable gate array IC chips 232. Computer 234 can be an appropriate personal computer (PC). Each time the transmitter of Fig. 1 is supplied with power, the gates of chips 232 are configured in accordance with the signals stored in flash random access memory 230. The configuration of the gates in chips 232 is changed at will, as are the filter coefficients by changing flash random access memory 230.

While there have been described and illustrated specific embodiments of the invention, it will be clear that variations in the details of the embodiments specifically illustrated and described may be made without departing from the true spirit and scope of the invention as defined in the appended claims.

BNSDOCID <WO 0178392A2 I >

WE CLAIM:

1. A method of preventing substantial interference between portions of first and second spectra that tend to overlap, the overlapping portion of the first spectrum being in a band of frequencies on a first side of a predetermined frequency of the first spectrum, the predetermined frequency being outside the second spectrum and within the first spectrum, comprising the steps of modifying the first spectrum prior to transmission thereof by transferring energy in frequencies on the first side of the predetermined frequency of the first spectrum to frequencies on a second side of the predetermined frequency so that all energy between a first frequency at a band edge of the first spectrum and a second frequency between the first frequency and the predetermined frequency is reduced substantially to zero, the frequencies on the second side being within the first spectrum, the transfer being such that a receiver designed to respond to the first spectrum causes the transferred energy to be shifted from the second side of the predetermined frequency to the first side of the predetermined frequency, and transmitting the modified spectrum, the amount of energy transferred being such that there is no substantial interference between the transmitted first and second spectra.

2. The method of claim 1 wherein the receiver includes a bandpass filter with a transition having a predetermined amplitude versus frequency response between a frequency at one band edge of the bandpass filter and a mid-range portion of the bandpass filter response, the predetermined frequency being displaced from a corner frequency between the mid-range portion and the transition toward the frequency at said one band edge, the energy transfer being such that (1) all energy between said first and second frequencies is transferred on a frequency to frequency basis to frequencies that are on the second side of the predetermined frequency and are equally displaced from the predetermined frequency, and (2) some energy between the second frequency and the predetermined frequency is transferred, on a frequency to frequency basis, to

frequencies that are on the second side of the predetermined frequency and are equally displaced from the predetermined frequency, the amount of the energy transferred from between the second frequency and the predetermined frequency being determined by the relative amplitudes of (a) an amplitude versus frequency weighting function that extends over a band of frequencies between the second frequency and the predetermined frequency, and (b) the amplitude versus frequency response of the receiver transition.

3. The method of claim 2 wherein the receiver bandpass filter has a flat mid-range amplitude versus frequency response and a root raised cosine transition between the corner frequency and the first frequency.

4. The method of claim 3 wherein the first spectrum is a DTV vestigial sideband spectrum.

5. The method of claim 4 wherein the predetermined frequency is a baseband pilot carrier frequency.

6. The method of claim 4 wherein the predetermined frequency is a Nyquist sampling frequency.

7. The method of claim 2 wherein the predetermined frequency is approximately half-way between the first and corner frequencies.

8. The method of claim 2 wherein the modified spectrum has a symmetrical amplitude versus frequency response with respect to its center frequency.

9. The method of claim 2 wherein the modified spectrum has an amplitude versus frequency response at its high frequency band edge including the transferred energy and an amplitude versus frequency response at its low frequency band edge that is substantially the same as the receiver transition.

173

10. The method of claim 2 wherein the modified spectrum has an amplitude versus frequency response at its low frequency band edge including the transferred energy and an amplitude versus frequency response at its high frequency band edge that is substantially the same as the receiver transition.

11. The method of claim 2 wherein the energy transfer is performed by multiplying a baseband signal containing information to be transmitted by sine and cosine representations to derive a pair of product signals, the sine and cosine representations being at a folding frequency, lowpass filtering replicas of the product signals, the lowpass filtering step including multiplying the replicas of the product signals by (a) a constant amplitude from DC to another frequency where the transition begins, (b) amplitudes greater than the constant amplitude for frequencies between the another frequency and a further frequency, (c) finite amplitudes less than the constant amplitude for frequencies greater than the further frequency and (d) zero for frequencies that are displaced from the another frequency by the same amount that said frequency at the band edge of the bandpass filter is displaced from the corner frequency and combining signals resulting from the lowpass filtering steps.

12. The method of claim 11 wherein the replicas on which the lowpass filtering steps are performed result directly from the multiplying operations to cause the spectrum resulting from the combining step to be symmetrical with respect to a frequency at the center of the first spectrum.

13. The method of claim 11 further including shifting the frequency of the spectra of the signals resulting from the lowpass filtering steps to cause the spectrum resulting from the combining step to be symmetrical with respect to a frequency having an exact numerical value.

14. The method of claim 11 further including performing second lowpass filtering steps on replicas of the product signals, the second lowpass

filtering steps being performed on the product signal replicas before the lowpass filtering operation defined in claim 11, the second lowpass filtering operations including multiplying the replicas of the product signal by (a) a constant amplitude from DC to the another frequency, and (b) the receiver transition amplitude versus frequency response for frequencies between the another frequency and a frequency removed from the another frequency by an amount equal to the displacement of the band edge of the receiver bandpass filter from the corner frequency, shifting the spectra resulting from the second lowpass filtering steps to cause the spectrum resulting from the combining step to be asymmetrical with respect to a frequency at the center of the first spectrum, the asymmetrical spectrum having the same amplitude versus frequency response at a first band edge as the receiver bandpass filter and to have an amplitude versus frequency response at a second band edge determined by the lowpass filtering operation of claim 11.

15. The method of claim 2 wherein the energy transfer includes lowpass filtering a signal resulting from a baseband signal containing information to be transmitted, the lowpass filtering step including multiplying the signal resulting from the baseband signal by (a) a constant amplitude from DC to another frequency where the transition begins, (b) amplitudes greater than the constant amplitude for frequencies between the another frequency and a further frequency, (c) finite amplitudes less than the constant amplitude for frequencies greater than the further frequency and (d) zero for frequencies that are displaced from the another frequency by the same amount that said frequency at the band edge of the bandpass filter is displaced from the corner frequency.

16. The method of claim 15 further including multiplying the signal resulting from the lowpass filtering step by sine and cosine representations to derive a pair of product signals, the sine and cosine representations being at a folding frequency, performing second lowpass filtering steps on replicas of the product signals, the second lowpass filtering steps being performed on the product

signal replicas before the lowpass filtering operation defined in claim 11, the second lowpass filtering operations multiplying the replicas of the product signal by (a) a constant amplitude from DC to the another frequency, and (b) the receiver transition amplitude versus frequency response for frequencies between the another frequency and a frequency removed from the another frequency by an amount equal to the displacement of the band edge of the receiver bandpass filter from the corner frequency,

17. The method of claim 16 further including multiplying the signals resulting from the second lowpass filtering step by cosine and sine representations of an IF to derive second product signals, and combining the second product signals to derive an asymmetrical spectrum that has the amplitude versus frequency response of the receiver transitions at its low frequency edge and the amplitude versus frequency response resulting from the lowpass filter operation of claim 15 at its high frequency edge.

18. The method of claim 15 further including performing I and Q channel Hilbert transform operations on a signal resulting from the baseband signal so that signals resulting from the I and Q channel Hilbert transform operations have the lowpass filtering operations of claim 15 applied to them, and combining replicas of the signals resulting from the I and Q channel Hilbert transform operations to derive the first spectrum.

19. The method of claim 18 wherein the combining step includes multiplying the signals resulting from the I and Q channel Hilbert transform operations with representations of cosine and sine waves at an IF to form product signals, and adding the product signals.

20. The method of claim 1 wherein the first spectrum is a DTV vestigial sideband spectrum and wherein the receiver has a bandpass filter with a flat mid-range amplitude versus frequency response and a root raised cosine transition between a corner frequency and the first frequency, the corner

frequency being the frequency where the receiver bandpass filter response changes from the mid-range amplitude to the root raised cosine transition.

21. The method of claim 20 wherein the predetermined frequency is a baseband pilot carrier frequency.

22. The method of claim 20 wherein the predetermined frequency is a Nyquist sampling frequency.

23. The method of claim 20 wherein the modified spectrum has a symmetrical amplitude versus frequency response with respect to its center frequency.

24. The method of claim 20 wherein the modified spectrum has an amplitude versus frequency response at its high frequency band edge including the transferred energy and an amplitude versus frequency response at its low frequency band edge that is substantially the same as the receiver transition.

25. The method of claim 20 wherein the modified spectrum has an amplitude versus frequency response at its low frequency band edge including the transferred energy and an amplitude versus frequency response at its high frequency band edge that is substantially the same as the receiver transition.

26. The method of claim 20 wherein the energy transfer includes lowpass filtering a signal resulting from a baseband DTV signal, the lowpass filtering step including multiplying the signal resulting from the baseband signal by (a) a constant amplitude from DC to another frequency where the transition begins, (b) amplitudes greater than the constant amplitude for frequencies between another frequency and a further frequency, (c) finite amplitudes less than the constant amplitude for frequencies greater than the further frequency and (d) zero for frequencies that are displaced from the another frequency by the same amount that said frequency at the band edge of the bandpass filter is displaced from the corner frequency.

27. The method of claim 26 further including multiplying the signal resulting from the lowpass filtering step by sine and cosine representations to derive a pair of product signals, the sine and cosine representations being at a folding frequency, performing second lowpass filtering steps on replicas of the product signals, the second lowpass filtering steps being performed on the product signal replicas before the lowpass filtering operation defined in claim 26, the second lowpass filtering operations multiplying the replicas of the product signal by (a) a constant amplitude from DC to the another frequency, and (b) the receiver transition amplitude versus frequency response for frequencies between the another frequency and a frequency removed from the another frequency by an amount equal to the displacement of the band edge of the receiver bandpass filter from the corner frequency,

28. The method of claim 27 further including multiplying the signals resulting from the second lowpass filtering step by cosine and sine representations of an IF to derive second product signals, and combining the second product signals to derive an asymmetrical spectrum that has the amplitude versus frequency response of the receiver transitions at its low frequency edge and the amplitude versus frequency response resulting from the lowpass filter operation of claim 26 at its high frequency edge.

29. The method of claim 26 further including performing I and Q channel Hilbert transform operations on a signal resulting from the baseband signal so that signals resulting from the I and Q channel Hilbert transform operations have the lowpass filtering operations of claim 26 applied to them, and combining replicas of the signals resulting from the I and Q channel Hilbert transform operations to derive the first spectrum.

30. The method of claim 29 wherein the combining step includes multiplying the signals resulting from the I and Q channel Hilbert transform

operations with representations of cosine and sine waves at an IF to form product signals, and adding the product signals.

31. A transmitter for a first spectrum having a portion that tends to overlap in an interfering manner with a neighboring, second spectrum, the overlapping portion being in a band of frequencies on a first side of a predetermined frequency, the predetermined frequency being outside the second spectrum and within the first spectrum, the transmitter comprising a filter for modifying the first spectrum prior to transmission, the filter being arranged for transferring energy in frequencies on the first side of the predetermined frequency of the first spectrum to frequencies on a second side of the predetermined frequency, the energy transfer causing substantial elimination from the first spectrum of all energy between a band edge of the first spectrum and another frequency between the band edge and the predetermined frequency, the frequencies on the second side being within the first spectrum, the transfer being such that receivers designed to respond to the first spectrum cause the transferred energy to be shifted from the second side of the predetermined frequency to the first side of the predetermined frequency, and an output device connected to be responsive to the modified spectrum.

32. The transmitter of claim 31 wherein the first spectrum is included in a signal source comprising a digital baseband television signal source and a carrier pilot inserted on the digital baseband television signal source, the filter including a digital vestigial sideband modulator connected to be responsive to the digital baseband television signal source and the carrier pilot inserted on the digital baseband television signal source.

33. The transmitter of claim 32 wherein the digital vestigial sideband modulator includes a digital filter including a Hilbert transform combined with a lowpass filter having an amplitude versus frequency response for causing the energy transfer.

34. The transmitter of claim 32 wherein the vestigial sideband modulator includes a digital folding frequency source for deriving orthogonal components at the predetermined frequency, a digital multiplier arrangement connected to be responsive to the orthogonal components and the digital signal source for deriving a pair of orthogonal product digital signal components, a digital lowpass filter arrangement connected to be responsive to the pair of orthogonal product digital signal components, an arrangement for combining the orthogonal digital signal components passed by the digital lowpass filter arrangement, the digital lowpass filter arrangement having a response and the digital signal components passed by the digital lowpass filter being combined so the energy transfer occurs and there is no substantial energy between the another frequency and the band edge of the first spectrum.

35. A memory for controlling an electromagnetic wave transmitter, the transmitter being arranged for transmitting a signal to a receiver having a predetermined response, the memory storing signals for preventing substantial interference between portions of first and second spectra that tend to overlap, the overlapping portion of the first spectrum being in a band of frequencies on a first side of a predetermined frequency of the first spectrum, the predetermined frequency being outside the second spectrum and within the first spectrum, the stored signals causing a modification of the first spectrum prior to transmission thereof by transferring energy in frequencies on the first side of the predetermined frequency of the first spectrum to frequencies on a second side of the predetermined frequency so that all energy between a first frequency at the band edge and a second frequency between the first frequency and the predetermined frequency is reduced substantially to zero, the frequencies on the second side being within the first spectrum, the transfer being such that a receiver designed to respond to the first spectrum causes the transferred energy to be shifted from the second side of the predetermined frequency to the first side of the predetermined frequency.

36. A transmitter for a first spectrum having a portion that tends to overlap in an interfering manner with a neighboring, second spectrum, the overlapping portion being in a band of frequencies on a first side of a predetermined frequency, the predetermined frequency being outside the second spectrum and within the first spectrum, the transmitter comprising a filter for modifying the first spectrum prior to transmission, the filter being adapted to be arranged for transferring energy in frequencies on the first side of the predetermined frequency of the first spectrum to frequencies on a second side of the predetermined frequency, the energy transfer causing substantial elimination from the first spectrum of all energy between a band edge of the first spectrum and another frequency between the band edge and the predetermined frequency, the frequencies on the second side being within the first spectrum, the transfer being such that receivers designed to respond to the first spectrum cause the transferred energy to be shifted from the second side of the predetermined frequency to the first side of the predetermined frequency, and an output device connected to be responsive to the modified spectrum.

37. The transmitter of claim 36 wherein the first spectrum is included in a signal source comprising a digital baseband television signal source and a carrier pilot inserted on the digital baseband television signal source, the filter including a digital vestigial sideband modulator connected to be responsive to the digital baseband television signal source and the carrier pilot inserted on the digital baseband television signal source.

lowpass filter having an amplitude versus frequency response for causing the energy transfer.

39. The transmitter of claim 37 wherein the vestigial sideband modulator includes a digital folding frequency source for deriving orthogonal components at the predetermined frequency, a digital multiplier arrangement connected to be responsive to the orthogonal components and the digital signal source for deriving a pair of orthogonal product digital signal components, a digital lowpass filter arrangement connected to be responsive to the pair of orthogonal product digital signal components, an arrangement for combining the orthogonal digital signal components passed by the digital lowpass filter arrangement, the digital lowpass filter arrangement having a response and the digital signal components passed by the digital lowpass filter being combined so the energy transfer occurs and there is no substantial energy between the another frequency and the band edge of the first spectrum.

40. The transmitter of claim 36 wherein the receivers include a bandpass filter with a transition having a predetermined amplitude versus frequency response between a frequency at one band edge of the bandpass filter and a mid-range portion of the bandpass filter response, the predetermined frequency being displaced from a corner frequency between the mid-range portion and the transition toward the frequency at said one band edge, the energy transfer being such that (1) all energy between said first and second frequencies is transferred on a frequency to frequency basis to frequencies that are on the second side of the predetermined frequency and are equally displaced from the

the predetermined frequency is transferred, on a frequency to frequency basis, to frequencies that are on the second side of the predetermined frequency and are equally displaced from the predetermined frequency, the amount of the energy transferred from between the second frequency and the predetermined frequency being determined by the relative amplitudes of (a) an amplitude versus frequency weighting function that extends over a band of frequencies between the second frequency and the predetermined frequency, and (b) the amplitude versus frequency response of the receiver transition.

41. The transmitter of claim 40 wherein the first spectrum is a DTV vestigial sideband spectrum.

42. The transmitter of claim 41 wherein the predetermined frequency is a baseband pilot carrier frequency.

43. The transmitter of claim 41 wherein the predetermined frequency is a Nyquist sampling frequency.

44. The transmitter of claim 40 wherein the predetermined frequency is approximately half-way between the first and corner frequencies.

45. The transmitter of claim 40 wherein the modified spectrum has a symmetrical amplitude versus frequency response with respect to its center frequency.

46. The transmitter of claim 40 wherein the modified spectrum has an amplitude versus frequency response at its high frequency band edge including the transferred energy and an amplitude versus frequency response at its low

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receivers.

47. The transmitter of claim 36 wherein the first spectrum is a DTV vestigial sideband spectrum and wherein the receivers have a bandpass filter with a flat mid-range amplitude versus frequency response and a root raised cosine transition between a corner frequency and the first frequency, the corner frequency being the frequency where the receiver bandpass filter response changes from the mid-range amplitude to the root raised cosine transition.

48. The transmitter of claim 47 wherein the predetermined frequency is a baseband pilot carrier frequency.

49. The transmitter of claim 47 wherein the predetermined frequency is a Nyquist sampling frequency.

50. The transmitter of claim 47 wherein the modified spectrum has a symmetrical amplitude versus frequency response with respect to its center frequency.

51. The transmitter of claim 47 wherein the modified spectrum has an amplitude versus frequency response at its high frequency band edge including the transferred energy and an amplitude versus frequency response at its low frequency band edge that is substantially the same as the receiver transition.

52. The transmitter of claim 47 wherein the modified spectrum has an amplitude versus frequency response at its low frequency band edge including the transferred energy and an amplitude versus frequency response at its high frequency band edge that is substantially the same as the receiver transition.

spectrum having a portion that tends to overlap in an interfering manner with a neighboring, second spectrum, the overlapping portion being in a band of frequencies on a first side of a predetermined frequency, the predetermined frequency being outside the second spectrum and within the first spectrum, the transmitter comprising a digital signal processing arrangement for modifying the first spectrum prior to transmission, a controller for causing the digital signal processing arrangement to transfer energy in frequencies on the first side of the predetermined frequency of the first spectrum to frequencies on a second side of the predetermined frequency, the energy transfer causing substantial elimination from the first spectrum of all energy between a band edge of the first spectrum and another frequency between the band edge and the predetermined frequency, the frequencies on the second side being within the first spectrum, the transfer being such that receivers designed to respond to the first spectrum cause the transferred energy to be shifted from the second side of the predetermined frequency to the first side of the predetermined frequency, and an output device connected to be responsive to the modified spectrum.

54. The transmitter of claim 53 wherein the digital television signal is a baseband signal and the first spectrum includes a carrier pilot inserted on the digital baseband television signal, the filter including a digital vestigial sideband modulator connected to be responsive to the digital baseband television signal source and the carrier pilot inserted on the digital baseband television signal source.

55. The transmitter of claim 54 wherein the digital vestigial sideband

lowpass filter having an amplitude versus frequency response for causing the energy transfer.

56. The transmitter of claim 54 wherein the vestigial sideband modulator includes a digital folding frequency source for deriving orthogonal components at the predetermined frequency, a digital multiplier arrangement connected to be responsive to the orthogonal components and the digital signal source for deriving a pair of orthogonal product digital signal components, a digital lowpass filter arrangement connected to be responsive to the pair of orthogonal product digital signal components, an arrangement for combining the orthogonal digital signal components passed by the digital lowpass filter arrangement, the digital lowpass filter arrangement having a response and the digital signal components passed by the digital lowpass filter being combined so the energy transfer occurs and there is no substantial energy between the another frequency and the band edge of the first spectrum.

57. The transmitter of claim 53 wherein the first spectrum is a vestigial sideband spectrum and wherein the receivers have a bandpass filter with a flat mid-range amplitude versus frequency response and a root raised cosine transition between a corner frequency and the first frequency, the corner frequency being the frequency where the receiver bandpass filter response changes from the mid-range amplitude to the root raised cosine transition.

58. The transmitter of claim 57 wherein the predetermined frequency is a baseband pilot carrier frequency.

59. The transmitter of claim 57 wherein the predetermined frequency is a Nyquist sampling frequency.

60. The transmitter of claim 57 wherein the modified spectrum has a symmetrical amplitude versus frequency response with respect to its center frequency.

61. The transmitter of claim 57 wherein the modified spectrum has an amplitude versus frequency response at its high frequency band edge including the transferred energy and an amplitude versus frequency response at its low frequency band edge that is substantially the same as the transition of the receivers.

62. The transmitter of claim 57 wherein the modified spectrum has an amplitude versus frequency response at its low frequency band edge including the transferred energy and an amplitude versus frequency response at its high frequency band edge that is substantially the same as the transition of the receivers.

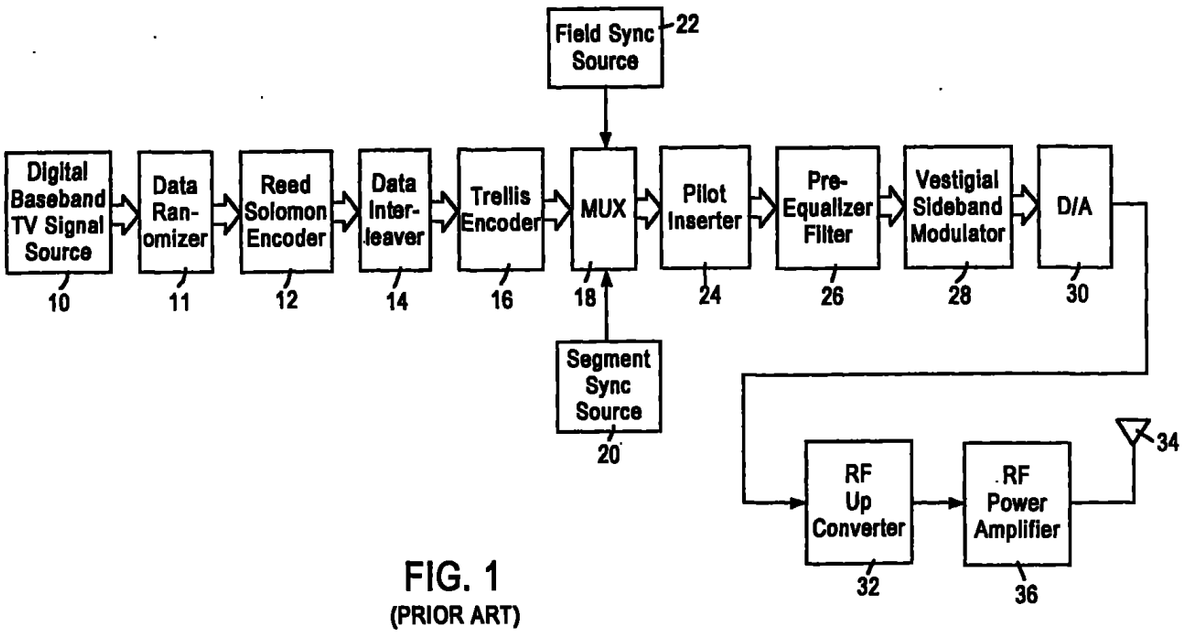


FIG. 1
(PRIOR ART)

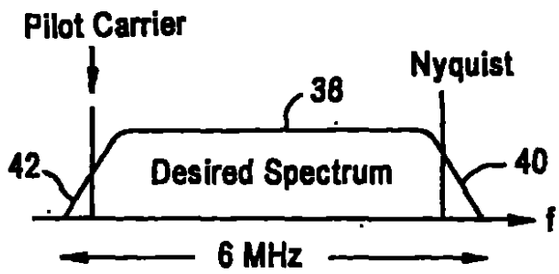


FIG. 2
(PRIOR ART)

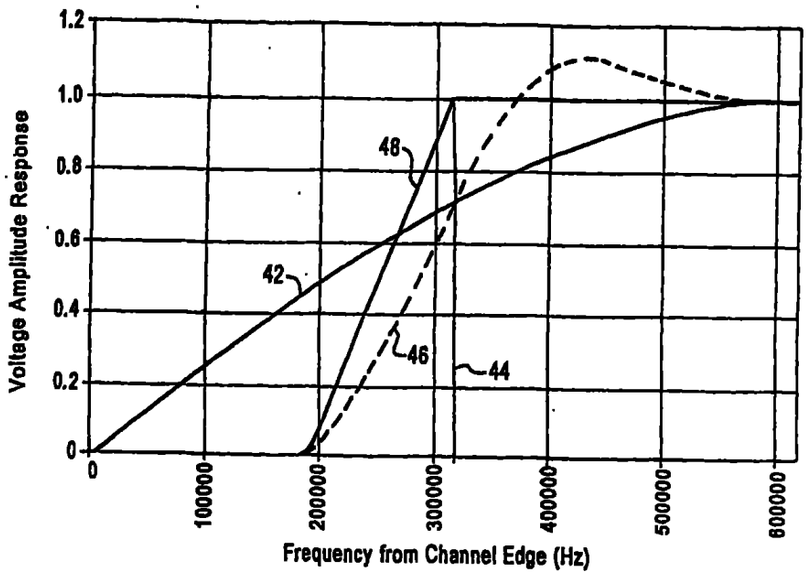

1.2
1.0
0.8
0.6
0.4
0.2
0
Voltage Amplitude Response
48
42
46
44
0
100000
200000
300000
400000
500000
600000
Frequency from Channel Edge (Hz)

FIG. 3

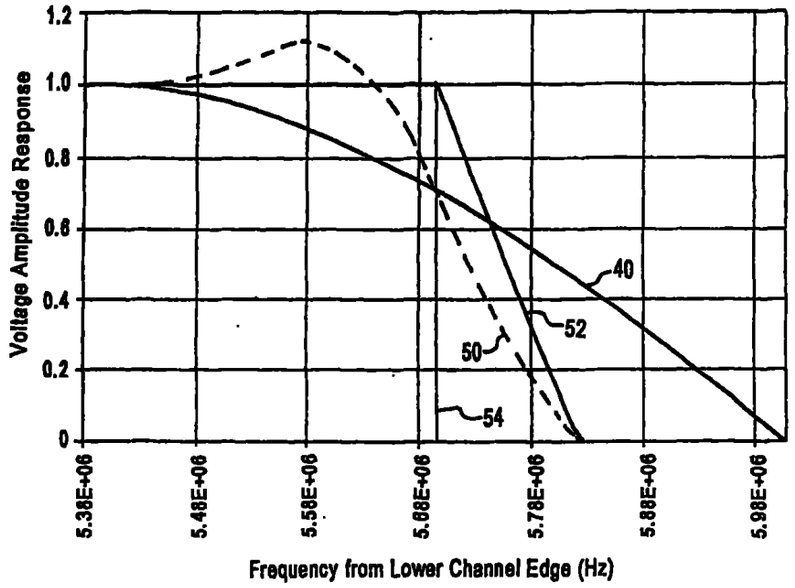

1.2
1.0
0.8
0.6
0.4
0.2
0
Voltage Amplitude Response
5.38E+06
5.48E+06
5.58E+06
5.68E+06
5.78E+06
5.88E+06
5.98E+06
Frequency from Lower Channel Edge (Hz)
40
52
50
54

FIG. 4

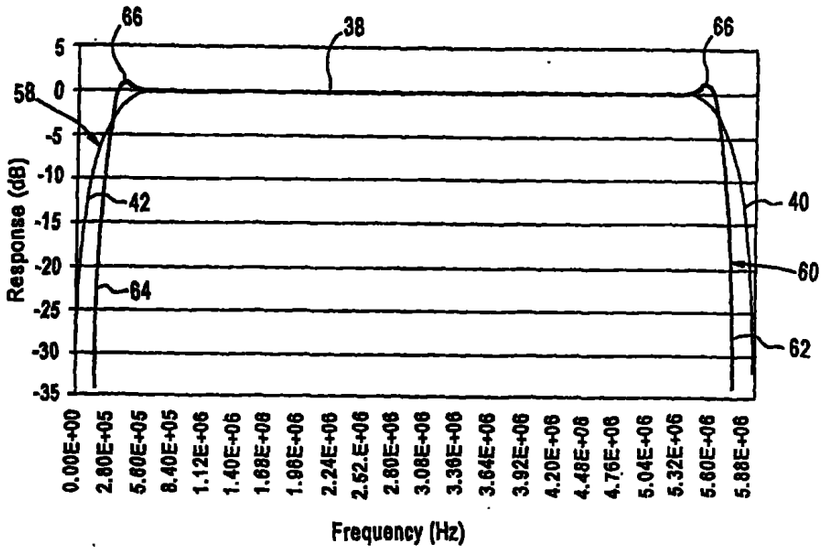

66
38
66
58
42
64
40
60
62
Response (dB)
5
0
-5
-10
-15
-20
-25
-30
-35
0.00E+00
2.80E+05
5.60E+05
8.40E+05
1.12E+06
1.40E+06
1.68E+06
1.96E+06
2.24E+06
2.52E+06
2.80E+06
3.08E+06
3.36E+06
3.64E+06
3.92E+06
4.20E+06
4.48E+06
4.76E+06
5.04E+06
5.32E+06
5.60E+06
5.88E+06
Frequency (Hz)

FIG. 5

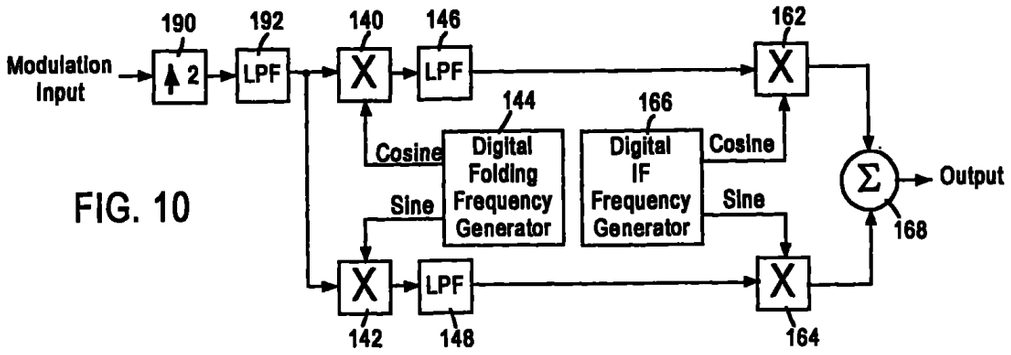

190
192
140
146
162
Modulation Input
2
LPF
X
LPF
X
144
166
Cosine
Digital Folding Frequency Generator
Digital IF Frequency Generator
Cosine
Σ
Output
Sine
Sine
168
X
LPF
X
142
148
164

FIG. 10

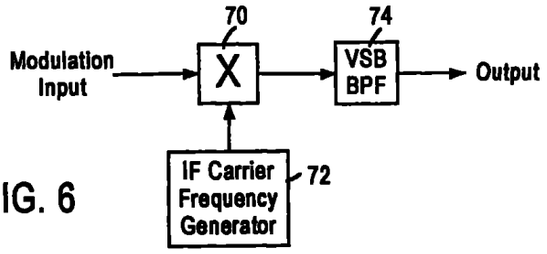

70
74
Modulation Input
X
VSB BPF
Output
IF Carrier Frequency Generator
72

FIG. 6

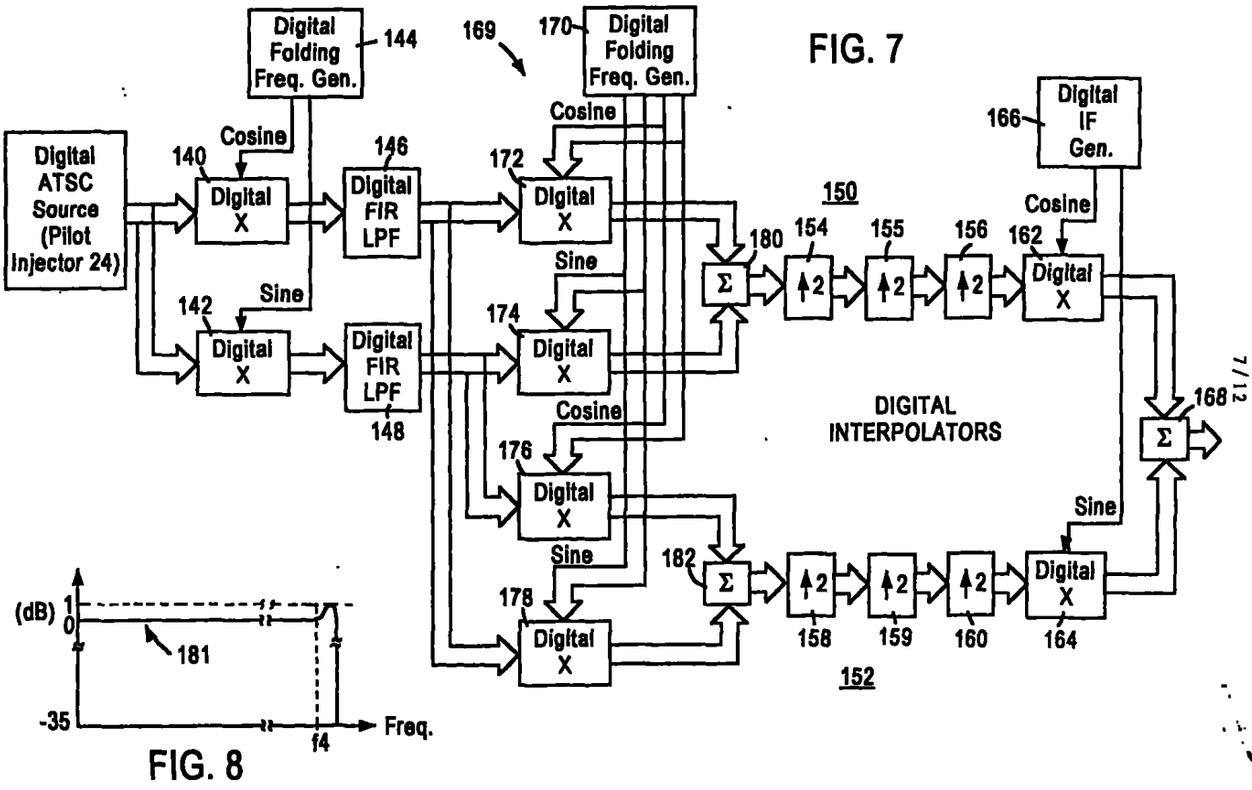

Digital Folding Freq. Gen.
144
169
170
Digital Folding Freq. Gen.
Digital ATSC Source (Pilot Injector 24)
Cosine
140
Digital X
146
Digital FIR LPF
172
Digital X
150
166
Digital IF Gen.
Cosine
180
154
155
156
162
Digital X
Sine
142
Digital X
Digital FIR LPF
148
174
Digital X
Cosine
DIGITAL INTERPOLATORS
168
176
Digital X
Sine
Sine
182
178
Digital X
158
159
160
164
Digital X
152

FIG. 7

(dB)
1
0
181
-35
f4
Freq.

FIG. 8

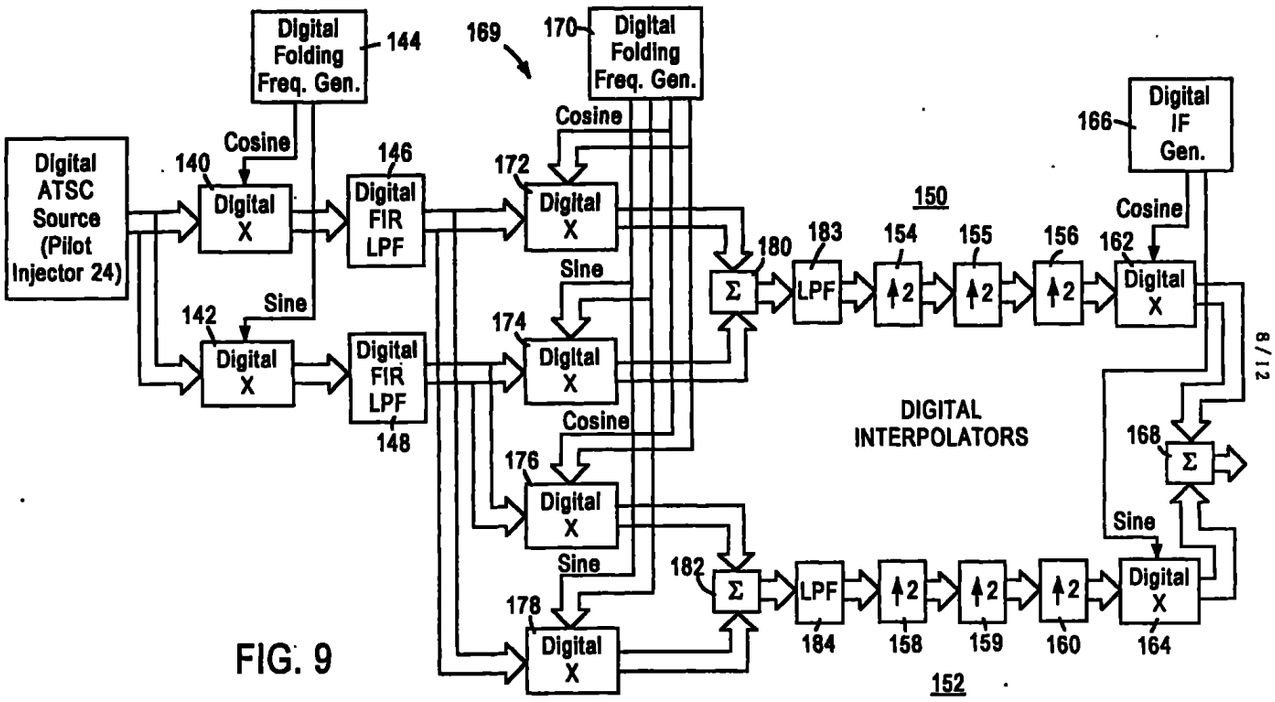

Digital Folding Freq. Gen.
144
169
170
Digital Folding Freq. Gen.
Cosine
166
Digital IF Gen.
Digital ATSC Source (Pilot Injector 24)
140
Cosine
Digital X
146
Digital FIR LPF
172
Digital X
150
Cosine
180
183
154
155
156
162
Σ
LPF
2
2
2
Digital X
Sine
142
Sine
Digital X
Digital FIR LPF
148
174
Digital X
DIGITAL INTERPOLATORS
168
Σ
Cosine
176
Digital X
Sine
Sine
182
178
Σ
LPF
2
2
2
Digital X
Digital X
184
158
159
160
164
152

FIG. 9

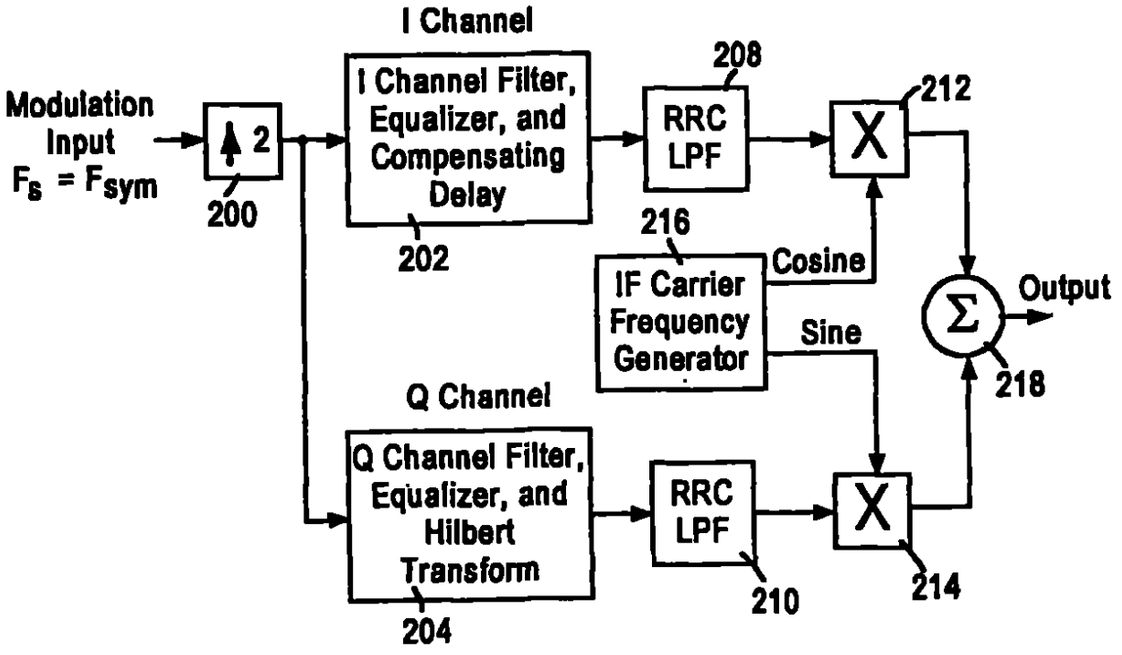

I Channel
Modulation Input
$F_S = F_{sym}$
2
200
I Channel Filter, Equalizer, and Compensating Delay
202
208
RRC LPF
212
X
216
IF Carrier Frequency Generator
Cosine
Sine
Output
Σ
218
Q Channel
Q Channel Filter, Equalizer, and Hilbert Transform
204
RRC LPF
210
X
214

FIG. 11

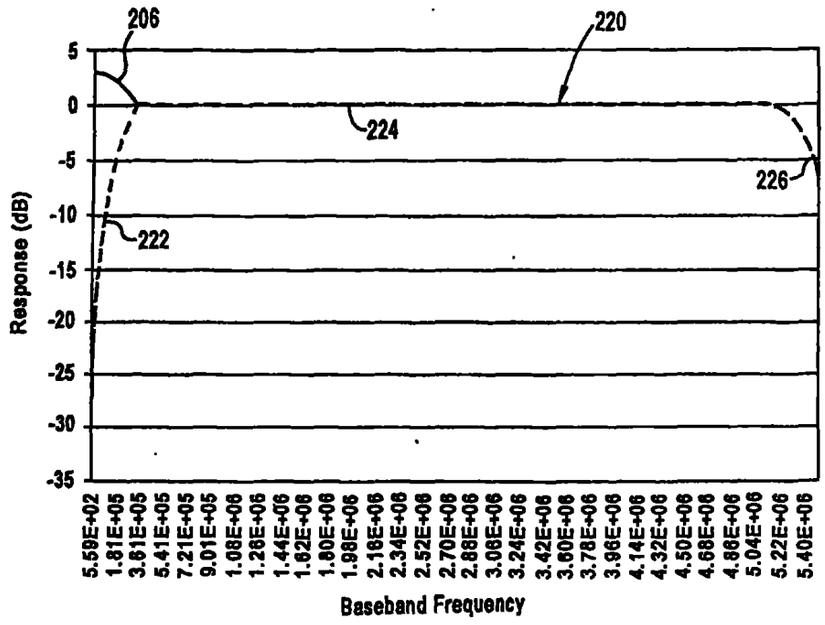

206
220
224
226
222
5
0
-5
-10
-15
-20
-25
-30
-35
Response (dB)
5.59E+02
1.81E+05
3.61E+05
5.41E+05
7.21E+05
9.01E+05
1.08E+06
1.26E+06
1.44E+06
1.62E+06
1.80E+06
1.98E+06
2.16E+06
2.34E+06
2.52E+06
2.70E+06
2.88E+06
3.06E+06
3.24E+06
3.42E+06
3.60E+06
3.78E+06
3.96E+06
4.14E+06
4.32E+06
4.50E+06
4.68E+06
4.86E+06
5.04E+06
5.22E+06
5.40E+06
Baseband Frequency

FIG. 12

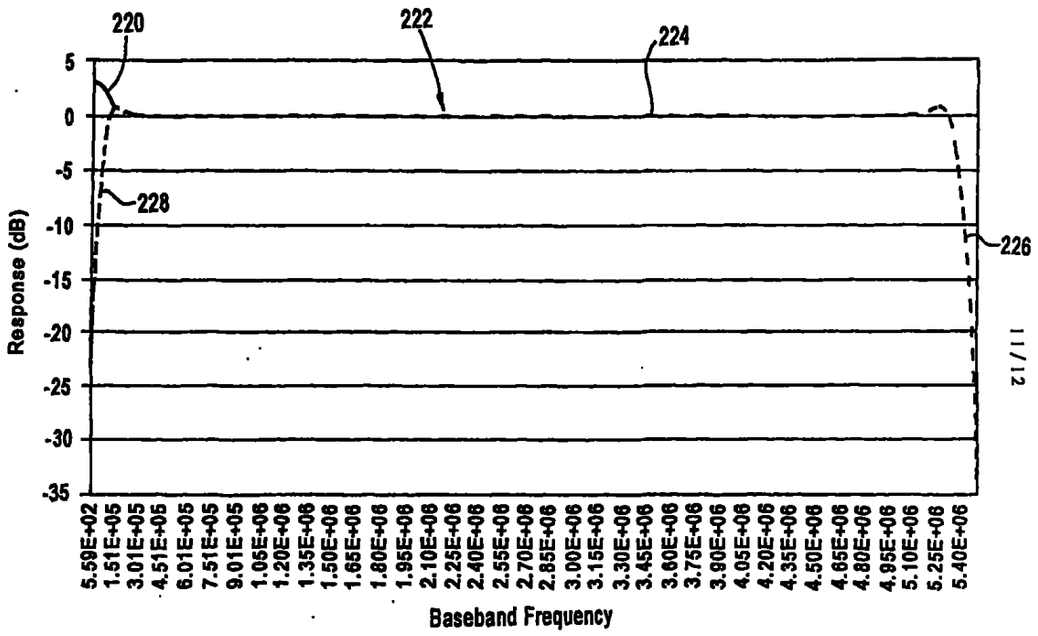

220
222
224
228
226
Response (dB)
5
0
-5
-10
-15
-20
-25
-30
-35
5.59E+02
1.51E+05
3.01E+05
4.51E+05
6.01E+05
7.51E+05
9.01E+05
1.05E+06
1.20E+06
1.35E+06
1.50E+06
1.65E+06
1.80E+06
1.95E+06
2.10E+06
2.25E+06
2.40E+06
2.55E+06
2.70E+06
2.85E+06
3.00E+06
3.15E+06
3.30E+06
3.45E+06
3.60E+06
3.75E+06
3.90E+06
4.05E+06
4.20E+06
4.35E+06
4.50E+06
4.65E+06
4.80E+06
4.95E+06
5.10E+06
5.25E+06
5.40E+06
Baseband Frequency

FIG. 13

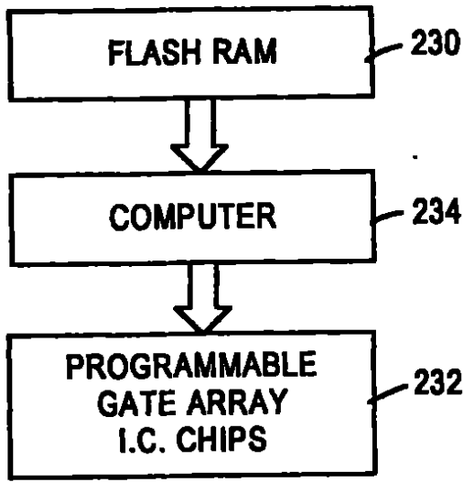

FLASH RAM
230
COMPUTER
234
PROGRAMMABLE
GATE ARRAY
I.C. CHIPS
232

FIG. 14

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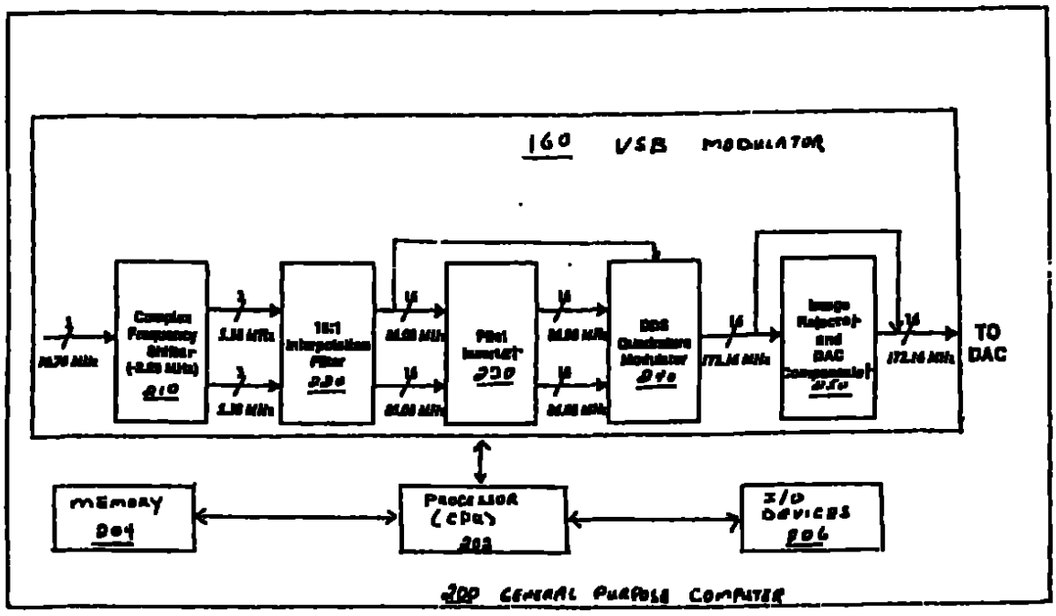
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(54) Title: DIRECT DIGITAL VESTIGIAL SIDEBAND (VSB) MODULATOR



(57) Abstract

A method and apparatus (160) for providing direct conversion of baseband digital symbols to a desired frequency (e.g., IF or RF) without the need for analog up conversion. The apparatus of the present invention is a VSB modulator comprising a complex frequency shifter (210), an interpolation filter (220), an optional pilot inserter (230), a DDS quadrature modulator (240) and an optional image rejector and DAC compensator (250). The architecture of the present VSB modulator can be implemented using digital integrated circuit technologies, thereby significantly reducing the cost of devices that require a VSB modulator.

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Direct Digital Vestigial Sideband (VSB) Modulator

This application claims the benefit of U.S. Provisional Application No. 60/120,569 filed on February 18, 1999, which is herein incorporated by reference.

The invention relates to a method and apparatus for modulating a vestigial sideband (VSB) signal.

BACKGROUND OF THE DISCLOSURE

With the ever increasing proliferation of digital information, various standards have been established to outline the format for implementing the transmission of such digital information. For example, digital data may be transmitted as a sub-carrier in the lower vestigial sideband (VSB) spectral region of a standard television signal, e.g., a National Television Standards Committee (NTSC) broadcast video signal.

Specifically, "8-VSB" and "16-VSB" are two radio frequency (RF) modulation formats that are utilized by the digital television standard (e.g., Advanced Television Systems Committee (ATSC)) to transmit digital information terrestrially or via a cable environment to consumers, respectively. An ATSC compliant transmitter will typically receive and modulate an encoded bitstream, e.g., an MPEG compliant bitstream, onto a RF channel and transmit the modulated signal to the consumers. With such wide acceptance of this digital technology, it is advantageous to obtain a transmitter having a low-cost digital VSB modulator that can be employed for high-definition television (HDTV) applications including remodulation of digital video directly to a typical television intermediate frequency (IF) or RF frequency.

Therefore, there is a need in the art for an apparatus and method for direct conversion of baseband digital symbols to a desired frequency (e.g., IF or RF) without the need for analog up conversion.

SUMMARY OF THE INVENTION

One embodiment of the present invention is a method and apparatus for providing direct conversion of baseband digital symbols to a desired frequency (e.g., IF or RF) without the need for analog up conversion.

Specifically, the apparatus of the present invention is a VSB modulator comprising a complex frequency shifter, an interpolation filter, an optional pilot inserter, a DDS quadrature modulator and an optional image rejector and DAC compensator.

The input to the complex frequency shifter can be an ATSC bitstream having undergone FEC processing. At the input of the VSB modulator, the symbols are frequency shifted to generate a complex signal and then decimated to provide two half-rate streams.

The frequency-shifted signal is then received by the interpolation filter, where the filter is a polyphase 16:1 interpolation filter having real coefficients. Namely, interpolation filter is a baseband interpolation filter. The polyphase 16:1 interpolation filter effectively generates 16 output symbols for every input symbol.

The pilot inserter adds a complex pilot signal at -2.69 MHz to the output of the interpolation filter. The pilot signal is a small DC shift that is applied as a helping signal. Namely, after frequency shaping the spectrum, the pilot signal is inserted at the left band edge. By performing pilot insertion after (rather than before) the interpolation filter, the datapath through the interpolation filter is maintained at 3 bits (rather than 16) so that the filter size is reduced. This is another advantage in reducing the overall cost of the modulator.

After pilot insertion, the data signal (a complex baseband signal) is applied to a quadrature modulator 240 that makes use of direct digital synthesizers (DDS) operating at 86.08 MHz. Two synthesizers are used to generate two real sinusoids that are 90 degrees out of phase from one another. Each of these sinusoids is multiplied by one of the input streams. Namely, the complex baseband signal to the quadrature modulator is digitally modulated up to some desired frequency, (e.g., intermediate frequency (IF) or RF). The desired frequency is selected within the DDS.

After multiplication, the two streams are multiplexed together to form a 176.16 MHz. output stream which contains the desired digital signal as well as images

due to the synthesizer aliases. This stream is then applied to the image rejector and DAC compensator 250, which is tasked with performing image rejection and compensation for spectral rolloff in digital-to-analog converters.

Thus, the present invention provides an apparatus and method for providing direct conversion of baseband digital symbols to a desired frequency (e.g., IF or RF) without the need for analog up conversion. Additionally, the architecture of the present VSB modulator can be implemented using digital integrated circuit technologies, thereby significantly reducing the cost of devices that require a VSB modulator.

BRIEF DESCRIPTION OF THE DRAWINGS

The teachings of the present invention can be readily understood by considering the following detailed description in conjunction with the accompanying drawings, in which:

FIG. 1 depicts a block diagram of a transmitter embodying the modulator of the present invention for direct conversion of baseband digital symbols to IF or RF without the need for analog up conversion;

FIG. 2 depicts a block diagram of a digital VSB modulator for direct conversion of baseband digital symbols to IF or RF without the need for analog up conversion; and

FIG. 3 is a flow diagram of a method for direct conversion of baseband digital symbols to IF or RF without the need for analog up conversion.

To facilitate understanding, identical reference numerals have been used, where possible, to designate identical elements that are common to the figures.

DETAILED DESCRIPTION

The invention will be described within the context of an ATSC television system. It will be apparent to one of ordinary skill in the art that the invention is also applicable to PAL and SECAM systems.

FIG. 1 depicts a block diagram of a transmitter 100 embodying the modulator of the present invention for direct conversion of baseband digital symbols to IF or RF without the need for analog up conversion. The transmitter 100 comprises a data randomizer 110, an encoder (e.g., Reed-Solomon encoder) 120, a data interleaver 130, an encoder (e.g., Trellis encoder) or a mapper 140, a multiplexer 150, a VSB modulator 160 and an RF up-converter 170.

The data randomizer 110 is tasked with randomizing the incoming data payload only, e.g., MPEG transport packets. Segment and Field Syncs are not randomized by the data randomizer 110. The randomization process is implemented to minimize concentration of RF energy at certain modulating frequency. Specifically, it is desirable to have the output VSB signal be as random as possible. The randomness ensures the transmitted signal frequency response will have a flat noise-like spectrum, thereby ensuring that the allotted channel space is used with maximum efficiency. Namely, if the data contains repetitious information, the recurring patterns may cause the RF energy content of the transmitted signal to congregate at certain discrete points in the frequency spectrum, thereby leaving voids at other frequencies. This scenario may create the undesirable condition of having the 6 MHz channel be overused in some parts and underused in other parts. The randomization pattern is a known pattern, such that the receiver can apply the reverse process to recover the proper data values. The present invention is not limited to a particular randomization process.

The encoder 120 is typically a block encoder, e.g., Reed-Solomon encoder, for applying a forward error correction (FEC) process to the incoming data signal. FEC processes are error correction schemes that are designed to correct bit errors that may occur during transmission, e.g., noise, multipath propagation, transmitter non-linearities, and signal fades. The present invention is not limited to a particular FEC process.

The data interleaver 130 is employed to scramble the sequential order of the data stream and to disperse the data packets throughout time, e.g., over a period of

time such as several milliseconds. The purpose of the data interleaver 130 is to disperse the data packets in order to minimize the transmitted signal's vulnerability to burst type interference. Namely, the data interleaver 130 assembles tiny fragments of scrambled data packets into new data packets, where the reconstituted data packets have the same length as the original data packets. Thus, a brief noise burst will only cause the lost of some of the data in the stream for any given period of time. ATSC defines the data interleaver employed in the VSB transmission system to be a 52 data segment (intersegment) convolutional byte interleaver and having a depth of approximately 1/6 of a data field (4 ms. deep). Again, the interleaving pattern is a known pattern, such that the receiver can apply the reverse process to recover the proper data. The present invention is not limited to a particular interleaving process.

The encoder 140 is typically a convolutional encoder, e.g., a trellis encoder for applying FEC. The trellis encoder employs a 2/3 rate (R=2/3) trellis code in accordance to ATSC. Specifically, each 2-bit word is compared to a past history of previous 2-bit words and a 3-bit binary code is generated to describe the transition from the previous 2-bit word to the current one. Namely, the 3-bit codes are substituted for the original 2-bit words and are transmitted as the symbols of the VSB. Again, the trellis code rate is defined by a particular standard, but the present invention is not limited to a particular trellis code rate.

The multiplexer 150 is tasked with inserting the segment sync and field sync. These sync signals are inserted after the randomization and error coding processes so as not to disturb the time and amplitude relationships of these signals.

In turn, the multiplexed signal is then modulated by the present novel VSB modulator 160 to produce a modulated signal. The modulated signal is then up-converted by the RF up-converter 170 to a proper RF frequency for transmission.

FIG. 2 depicts a block diagram of a digital VSB modulator 160 for direct conversion of baseband digital symbols to a desired frequency (e.g., IF or RF) without the need for analog up conversion. FIG. 3 is a flow diagram of a corresponding method 300 for the direct conversion of baseband digital symbols to a desired

frequency (e.g., IF or RF) without the need for analog up conversion. The reader will best understand the invention by simultaneously consulting both FIGs. 2 and 3.

The VSB modulator 160 performs modulation and filter shaping on the incoming signal. Specifically, the architecture of the present VSB modulator provides an important advantage of being a low-cost implementation of an all digital VSB modulator that can be employed for HDTV applications including remodulation of digital video to a typical television IF or RF frequency. This architecture can be implemented using digital integrated circuit technologies for direct conversion of baseband digital symbols to IF or RF without the need for analog up conversion.

Specifically, the VSB modulator 160 comprises a complex frequency shifter 210, an interpolation filter 220, a pilot inserter 230, a DDS quadrature modulator 240 and an image rejector and DAC compensator 250. As discussed below, the pilot inserter 230 and the image rejector and DAC compensator 250 can be deemed as optional to the extent that the functions provided by these modules can be omitted and/or performed outside of the present VSB modulator 160.

The input to the complex frequency shifter 210 can be an ATSC bitstream having undergone FEC processing. For example, the inputs to the VSB modulator 160 are 3-bit (8-level) symbols or 4-bit (16-level) symbols produced by a symbol mapper as described by the ATSC specification for digital video transmission. The mapping is typically applied to a digital video bitstream after it has been modified using forward error correction techniques, and the symbols are generated at a 10.76 MHz. rate.

At the input of the VSB modulator 160, the symbols are frequency shifted (step 310 of FIG. 3) to generate a complex signal and then decimated (step 320 of FIG. 3) to provide two half-rate streams. This operation can be performed very inexpensively using simple inverters and a demultiplexer. Namely, the input signal is filter-shaped by being frequency shifted to the "left" by −2.69 MHz. to allow filtering at baseband. The decimation operation splits the input stream (at 10.76 MHz.) into two streams (the I and Q components of a complex signal), each at only 5.38 MHz.

This decimation allows further processing to be performed more easily at half the required processing rate. Although the present complex frequency shifter 210 is described as performing two functions: frequency shifting and decimation, it should be understood that these two functions can be implemented within a single module or separate modules.

The frequency-shifted signal is then received by the interpolation filter 220 (step 330 of FIG. 3), where the filter is a polyphase 16:1 interpolation filter having real coefficients. Namely, interpolation filter 220 is a baseband interpolation filter. The polyphase 16:1 interpolation filter effectively generates 16 output symbols for every input symbol. The increased output rate or sampling rate allows for subsequent sampling condition where a multiplication by a high frequency sinusoidal is necessary (e.g., step 350 of FIG. 3). The advantage of performing this sampling rate conversion at this point is because the interpolation filter 220 is relatively inexpensive.

It should be noted that the coefficients of the interpolation filter 220 are specified so that the filter performs a root-raised cosine spectral shaping of the input signal with the rolloff factor set to 0.115 as specified by the ATSC. The bandwidth of this filter is only 2.69 MHz. Because of the decimation performed in the previous block, this filter only needs to operate at 86.08 MHz (16 x 5.38) rather than at twice the rate. The filter can be implemented using significant resource sharing or canonical signed digital techniques to control the implemented area of the filter. Typical coefficient sizes are on the order of 12-16 bits. The high interpolation ratio is important since the present VSB modulator is designed to perform the modulation process digitally instead of analog, thereby requiring a high sampling rate.

The pilot inserter 230 adds (step 340 of FIG. 3) a complex pilot signal at -2.69 MHz to the output (interpolated signals) of the interpolation filter 220. The pilot signal is a small DC shift that is applied as a helping signal. Namely, after frequency shaping the spectrum, the pilot signal is inserted at the left band edge. This causes a small residual carrier to appear at the zero frequency point of the resulting

modulated spectrum, thereby providing the receiver with a easily identifiable point to lock onto that is independent of the data being transmitted.

It should be noted that pilot insertion is typically performed before frequency shaping as disclosed in the ATSC standard. However, by performing pilot insertion after (rather than before) the interpolation filter, the datapath through the interpolation filter is maintained at 3 bits (rather than 16) so that the filter size is reduced. This again is another advantage in reducing the overall cost of the modulator. However, the pilot inserter 230 is optional to the extent that its function can be implemented prior to the VSB modulator as indicated by the ATSC standard. In such situation, a more complex (i.e., a more expensive) interpolation filter 220 is required since the interpolation filter cannot be maintained at 3 bits. Nevertheless, the VSB modulator is still capable of direct conversion of baseband digital symbols to IF or RF without the need for analog up conversion.

After pilot insertion, the data signal (a complex baseband signal) is applied (step 350 of FIG. 3) to a quadrature modulator 240 that makes use of direct digital synthesizers (DDS) operating at 86.08 MHz. Two synthesizers are used to generate two real sinusoids that are 90 degrees out of phase from one another. Each of these sinusoids is multiplied by one of the input streams. Namely, the complex baseband signal to the quadrature modulator 240 needs to be modulated up to some desired intermediate frequency (IF) or RF. The frequency of the sinusoids are typical television IF and RF frequencies, for example 43.75 MHz. (IF), 58 MHz. (Channel 3), or 64 MHz. (Channel 4). It should be noted that 58 MHz., or 64 MHz. are equivalent to typical VCR output frequencies, thereby allowing the present invention to be implemented in a digital VCR. Thus, the complex signals into the DDS quadrature modulator 240 is up-converted into a real signal at a desired frequency (IF or RF) at the output of the DDS quadrature modulator. The desired frequency is selected within the DDS. In essence, sub-sampling (e.g., sampling rate of 86.08 MHz.) is implemented to attain the desired frequency of 43.75 MHz., 58 MHz., 64 MHz. and so on. It should be noted that the desired frequency should be selected in a

manner that will not cause the images to overlap the desired frequency spectrum. For example, the sampling rate of 86.08 MHz. is chosen such that the desired frequency of 43.75 MHz., 58 MHz., and 64 MHz. can be attained without having distortion caused by overlap of images on the desired frequency spectrum.

However, since the synthesizer rate is only 86.08 MHz., aliases or images of the desired frequencies will appear at the output of the synthesizers and will be compensated by using post processing. Namely, the sampling rate is not twice the desired frequency, thereby failing to satisfy the Nyquist criterion and artifacts may result.

After multiplication, the two streams are multiplexed together to form a 176.16 MHz. output stream which contains the desired digital signal as well as images due to the synthesizer aliases. This stream is then applied to the image rejector and DAC compensator 250, which is tasked with performing image rejection and compensation for spectral rolloff in digital-to-analog converters (step 360 of FIG. 3).

In one embodiment, the image rejector and DAC compensator 250 comprises a real FIR filter or COMB filter for image rejection (or suppression). For image rejection, the filter coefficients will need to be programmable since the images will appear at different frequencies for each of the desired IF or RF outputs.

It should be understood that the image rejector and DAC compensator 250 is an optional module within the VSB modulator 160. Specifically, a SAW filter is typically implemented outside of the VSB modulator (e.g., after the DAC) that is tasked for filtering some of the images. However, if additional image rejection beyond that of the SAW filter is required, then the image rejector and DAC compensator 250 can be implemented to provide additional image rejection. For example, a notch can be implemented using a COMB filter in the image rejector and DAC compensator 250 at the image location to provide further attenuation on the image, thereby assisting subsequent analog filtering and reducing the complexity and cost of analog filtering. Thus, again the present VSB can be implemented to reduce the overall cost of the VSB transmitter.

Additionally, DAC generally generates "zero order hold", where a signal is held until the next sample, thereby generating a stair-case like response. The impact of this zero order hold, looking at the frequency domain, is similar to a low-pass filtering effect at the output of the DAC. Namely, distortion is introduced due to this low-pass roll-off. Thus, to pre-empt or compensate the effect of the DAC, the image rejector and DAC compensator 250 may apply a filter having a high-pass type of response to the oversampled input signal of 172.16 MHz. In sum, the effect of the high pass response caused by the image rejector and DAC compensator 250 when combined with the low-pass filtering effect of the DAC will produce the desired spectrum.

It should be noted that roll-off in the spectrum can generally be addressed by the receiver, thereby making the image rejector and DAC compensator 250 an optional module. However, if the present VSB modulator is implemented at a broadcast transmitter, it may be desirable to implement the image rejector and DAC compensator 250 to produce an output signal that is as clean as possible. Thus, for low performance system, the image rejector and DAC compensator 250 can be optionally omitted to further reduce cost of the VSB modulator.

The output of the VSB modulator can be applied to a DAC (not shown) operating at 172.16 MHz for direct generation of the IF or RF modulated analog signal. A post-processing analog bandpass filter can be applied after the DAC for further rejection of the out-of-band images and quantization noise.

Typically, two stages of up-conversions are necessary, where a baseband signal is first up-converted to an intermediate frequency, and is then up-converted again to a broadcast radio frequency. Thus, a prior art implementation will typically apply a DAC to the baseband signal and then apply a first analog up-conversion to attain the intermediate frequency and then apply a second analog up-conversion to attain the broadcast radio frequency.

In contrast, the present novel direct digital VSB modulator 160 up-converts the baseband signal digitally to the desired frequency, e.g., IF, thereby avoiding one

analog up-conversion. The DAC is then applied to the IF signal and then one analog up-conversion is performed to the attain the RF signal for broadcast.

FIG. 2 also depicts one embodiment of the present invention where the VSB modulator 160 is implemented within a general purpose computer. In one embodiment, the general purpose computer 200 comprises a VSB modulator 160, a processor (CPU) 202, a memory 204, e.g., random access memory (RAM), and various input/output devices 206, (e.g., a keyboard, a mouse, an audio recorder, a camera, a camcorder, a video monitor, any number of imaging devices or storage devices, including but not limited to, a tape drive, a floppy drive, a hard disk drive or a compact disk drive).

It should be understood that the VSB modulator 160 can be a physical device that is coupled to the CPU 202 through a communication channel. Alternatively, the VSB modulator 160 can be represented by one or more software applications (or even a combination of software and hardware, e.g., using application specific integrated circuits (ASIC)), where the software is loaded from a storage medium, (e.g., a magnetic or optical drive or diskette) and operated by the CPU in the memory 204 of the computer. As such, the VSB modulator 160 (including associated data structures) of the present invention can be stored on a computer readable medium, e.g., RAM memory, magnetic or optical drive or diskette and the like.

It should be understood that although the above disclosure provides a listing of various frequency values, these values are only approximate values and may vary within a practical application environment. In fact, some of these frequency values may vary significantly due to a particular implementation.

Although various embodiments which incorporate the teachings of the present invention have been shown and described in detail herein, those skilled in the art can readily devise many other varied embodiments that still incorporate these teachings.

What is claimed is:

1. Apparatus (160) for up-converting an input signal to a desired frequency, said apparatus comprising:

a shifter (210), for applying a frequency shift and decimation to the input signal to generate at least one frequency shifted signal;

an interpolation filter (220), coupled to said shifter, for interpolating said at least one frequency shifted signal to produce at least one interpolated signal; and

a direct digital synthesizers (DDS) modulator (240), coupled to said interpolation filter, for up-converting said at least one interpolated signal to the desired frequency.

2. The apparatus (160) of claim 1, further comprising:

a pilot inserter (230), coupled to said interpolation filter, for inserting a pilot signal onto said at least one interpolated signal.

3. The apparatus (160) of claim 1, further comprising:

an image rejector and digital-to-analog converter (DAC) compensator (250), coupled to said direct digital synthesizers (DDS) modulator, for rejecting images and for performing digital-to-analog compensation.

4. The apparatus (160) of claim 1, wherein said interpolation filter is a 16:1 interpolation filter.

5. The apparatus (160) of claim 1, wherein said input signal is a baseband signal.

6. Method for up-converting an input signal to a desired frequency, said method comprising the steps of:

(a) applying a frequency shift and decimation to the input signal to generate at least one frequency shifted signal;

(b) interpolating said at least one frequency shifted signal to produce at least one interpolated signal; and

(c) applying a direct digital synthesizers (DDS) modulator, to up-convert said at least one interpolated signal to the desired frequency.

7. The method of claim 6, further comprising the step of:

(b') inserting a pilot signal onto said at least one interpolated signal prior to said applying step (c).

8. The method of claim 6, further comprising the step of:

(d) rejecting images and performing digital-to-analog compensation on said at least one up-converted signal from step (c).

9. A computer-readable medium (206) having stored thereon a plurality of instructions, the plurality of instructions including instructions which, when executed by a processor, cause the processor to perform the steps comprising of:

(a) applying a frequency shift and decimation to the input signal to generate at least one frequency shifted signal;

(b) interpolating said at least one frequency shifted signal to produce at least one interpolated signal; and

(c) applying a direct digital synthesizers (DDS) modulator, to up-convert said at least one interpolated signal to the desired frequency.

10. A transmitter (100) for transmitting an input signal, said transmitter comprising:

a data randomizer (110) for randomizing said input signal to generate a randomized input signal;

a first encoder (120), coupled to said data randomizer, for applying block coding to said randomized input signal to generate a block coded input signal;

a data interleaver (130), coupled to said first encoder, for interleaving said block coded input signal to generate an interleaved input signal;

a second encoder (140), coupled to said data interleaver, for applying convolutional coding to said interleaved input signal to generate a convolutional coded input signal;

a multiplexer (150), coupled to said second encoder, for multiplexing said convolutional coded input signal with at least one sync signal to generate a multiplexed input signal;

a modulator (160), coupled to said multiplexer, said modulator comprising:

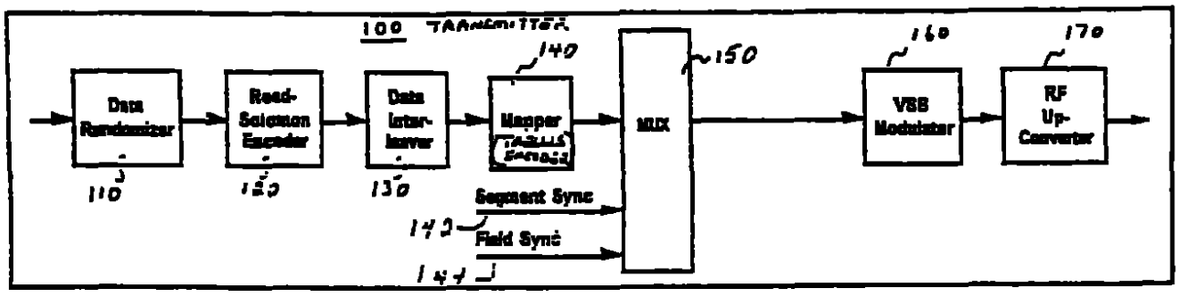
a shifter (210), for applying a frequency shift and decimation to the multiplexed input signal to generate at least one frequency shifted signal;

an interpolation filter (220), coupled to said shifter, for interpolating said at least one frequency shifted signal to produce at least one interpolated signal; and

a direct digital synthesizers (DDS) modulator (240), coupled to said interpolation filter, for up-converting said at least one interpolated signal to the desired frequency to generate a modulated signal; and

a up-converter (170), coupled to said modulator, for up-converting said modulated signal.

ENSDOCID <WO _ 0049710A1_I_>


100
Data Randomizer
110
Reed-Solomon Encoder
120
Data Interleaver
130
Mapper
140
MUX
150
VSB Modulator
160
RF Up-Converter
170
Segment Sync
142
Field Sync
144

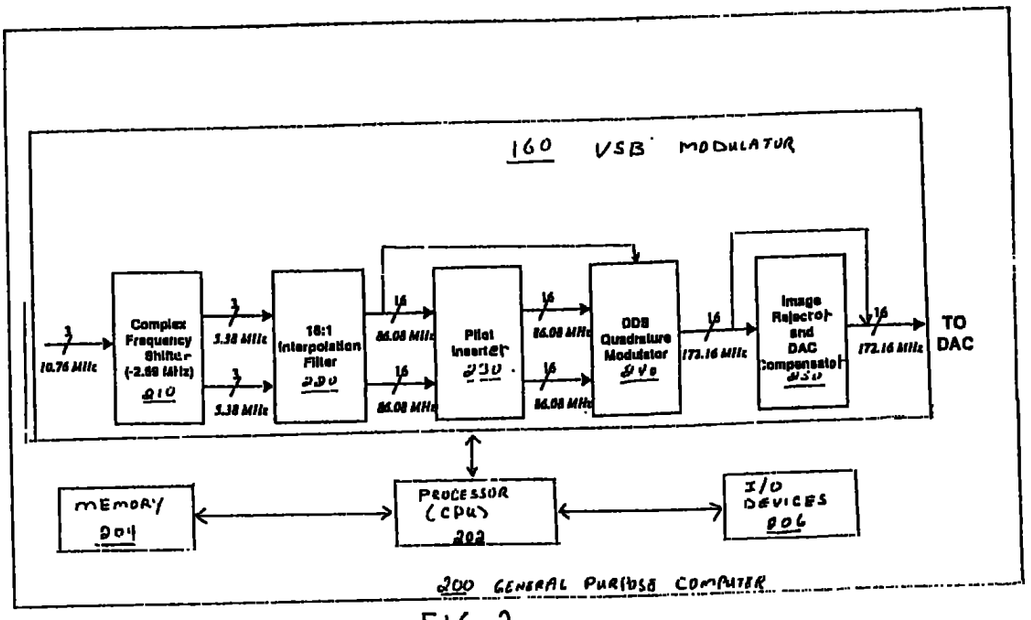

160 VSB MODULATOR
Complex Frequency Shifter (-2.69 MHz) 210
10.76 MHz
5.38 MHz
5.38 MHz
16:1 Interpolation Filter 220
86.08 MHz
86.08 MHz
Pilot Inserter 230
86.08 MHz
86.08 MHz
QDS Quadrature Modulator 240
173.16 MHz
Image Rejector and DAC Compensator 250
172.16 MHz
TO DAC
MEMORY 204
PROCESSOR (CPU) 202
I/O DEVICES 206
200 GENERAL PURPOSE COMPUTER

FIG. 2

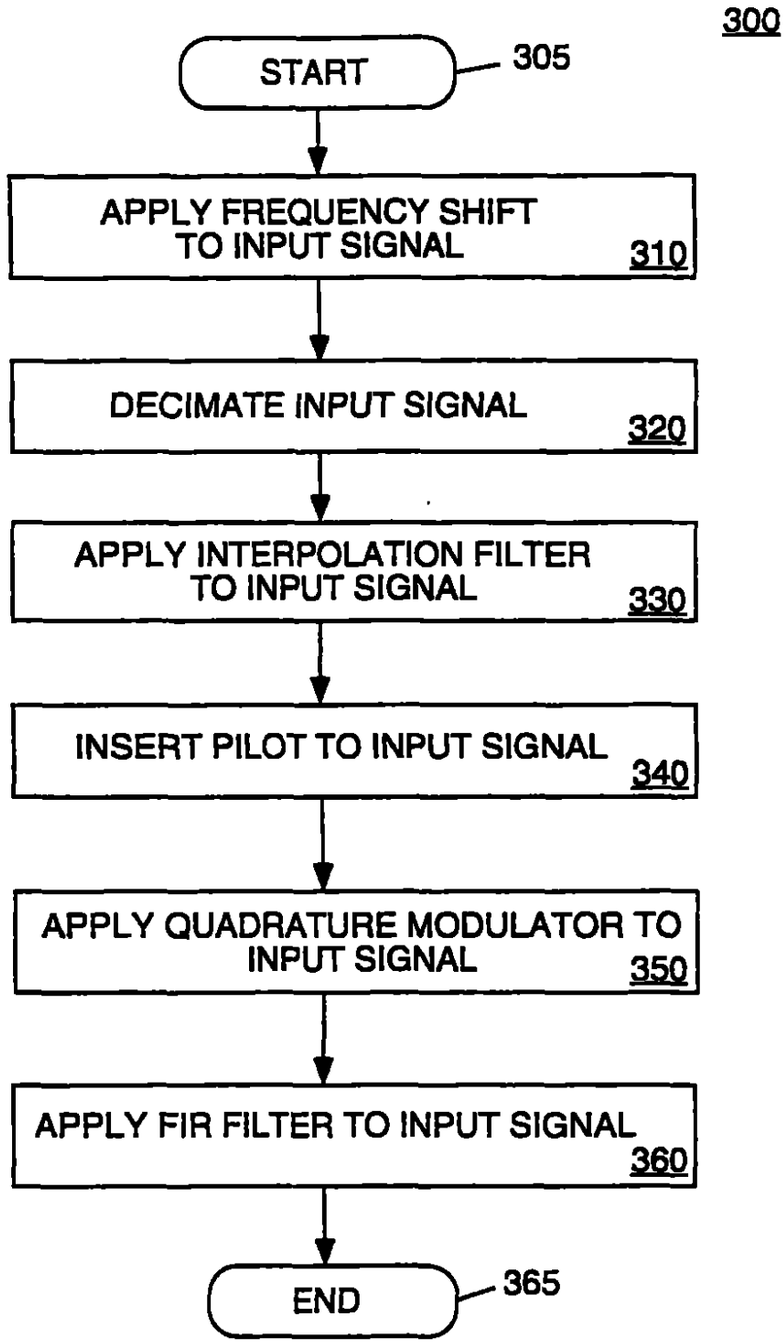

300
START
305
APPLY FREQUENCY SHIFT TO INPUT SIGNAL
310
DECIMATE INPUT SIGNAL
320
APPLY INTERPOLATION FILTER TO INPUT SIGNAL
330
INSERT PILOT TO INPUT SIGNAL
340
APPLY QUADRATURE MODULATOR TO INPUT SIGNAL
350
APPLY FIR FILTER TO INPUT SIGNAL
360
END
365

FIG. 3

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US00/04080

A. CLASSIFICATION OF SUBJECT MATTER

IPC(7) :H03C 1/52, 1/60; H04N 5/40
US CL : 375/ 270, 301; 332/170; 348/724; 455/ 109

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 375/ 257, 270, 301, 321, 346; 332/170; 348/607, 608, 724; 455/ 109

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

EAST, IEL

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X,P	US 5,949,878 A (BURDGE et al.) 07 September 1999, Fig. 1A	1, 5, 6, 9
Y,P		4
Y	US 4,974,236 A (GURCAN et al.) 17 November 1990, Fig. 2	1, 4-6, 9, 10
Y	US 5,412,352 A (GRAHAM) 02 May 1995, Fig. 4	1, 4-6, 9, 10
Y	US 5,508,752 A (KIM et al.) 16 April 1996, Fig. 1	10
A	US 5,801,595 A (DAVIS et al.) 01 September 1998, whole document	1-9

[X] Further documents are listed in the continuation of Box C. [ ] See patent family annex.

*	Special categories of cited documents:	"T"	later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"A"	document defining the general state of the art which is not considered to be of particular relevance	"X"	document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"E"	earlier document published on or after the international filing date	"Y"	document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"L"	document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)	"&"	document member of the same patent family
"O"	document referring to an oral disclosure, use, exhibition or other means		
"P"	document published prior to the international filing date but later than the priority date claimed		

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Form PCT/ISA/210 (second sheet) (July 1998) *

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INTERNATI NAL SEARCH REPORT

International application No.
PCT/US00/04080

C (Continuation). DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US 5,512,957 A (HULYALKAR) 30 April 1996, Fig. 1	1-10

BNSDOCID: <WO 0049710A1 I >

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(54) Device and method for digital vestigial sideband modulation

(57) A digital signal processing device and method for vestigial sideband (VSB) modulation of a digital signal. A binary digital signal d_l is provided to a baseband shaping filter which samples the signal to provide a baseband signal which has real and imaginary portions $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$, where

$$S_{bb}(kT) = \sum_l (d_l + b)\left(\cos(\frac{\Pi}{2}l) - j\sin(\frac{\Pi}{2}l)\right) h_0((k-l)T)$$

The VSB signal is provided with perfect sideband suppression without using analog filters.

EP 0 853 409 A2

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Description

The present invention is directed to digital signal processing, and specifically to a device and method for modulating a digital television (DTV) signal as a vestigial sideband (VSB) signal occupying a standard television channel.

VSB modulation is proposed for the U. S. A. standard for future broadcast of DTV. It is proposed of a particular type of VSB modulation, known as 8-VSB, which maps 3-bit groups of coded digital data onto one of eight amplitude levels for transmission.

One of the problems encountered in VSB modulation is sideband suppression. Various VSB modulation methods have been proposed which use analog filters (e.g., SAW filters) for sideband suppression. However, analog filters require more bandwidth than is desirable.

The present invention uses digital signal processing to achieve perfect sideband suppression without analog filters. The processing is characterized by requiring minimal bandwidth in the shaping filter, a characteristic which provides cascaded improvements. For example, a DTV digital signal can be represented over a wider bandwidth for a given sampling rate, which provides the advantage of allowing adaptive corrections in the digital domain to compensate for linear and nonlinear characteristics of high power television transmitters. Further, the present invention allows the control of out-of-band splatter and spectral sidelobes from the television transmitter, thereby reducing interference with adjacent channels. By way of further example, it is known that third order intermodulation products must be included in the signal for predistortion, and the digital processing method and device herein controls the representation of such products within three times the main spectral lobe width of the television signal.

Analog devices are typically manually adjusted, do not track time-varying transmitter distortions, and cannot take advantage of adaptive features allowed by the digital system herein.

The device and method of vestigial sideband modulation herein may include the provision of an input digital signal (not limited to a DTV signal) at a rate 1/T to a multiplier which downconverts the input signal by multiplying it by $e^{-j(2\Pi/4T)t}$. The downconverted signal may thereafter be provided to a symmetrical shaping filter which Nyquist filters the downconverted signal and has an impulse response of $h_0(t)$, and the filtered signal may then be upconverted so that it is centered on a carrier frequency, ω_0, in the vestigial sideband of the input signal by multiplying the filtered signal by

$$e^{j(\omega_0 + (2\Pi/4T))t}$$

to provide a vestigial sideband modulated output signal.

An object of the present invention is to provide a device and method for VSB modulating a digital signal which uses digital signal processing to avoid the use of analog filters, and to provide a digital device and digital signal processing method for VSB modulation in which a digital signal is provided to a baseband shaping filter which provides a baseband signal with real and imaginary portions $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$, where

$$S_{bb}(kT) = \sum_i (d_i + b)\left(\cos(\frac{\Pi}{2}i) - j\sin(\frac{\Pi}{2}i)\right) h_0((k-i)T)$$

and in which the real and imaginary portions are interpolated so that a vestigial sideband modulated output signal includes third order intermodulation distortion products of the signal.

Another object is to provide a baseband filter for VSB modulation of a DTV signal which has,

a multiplier for multiplying an arriving television signal by $(\cos(\Pi i/2) - \sin(\Pi i/2))$,
a transversal filter for providing separate sums of even and odd coefficient terms,
an adder for adding a DC offset term to the sum of even coefficient terms, and
a multiplexer for multiplexing the sum of odd coefficient terms with the sum of even coefficient terms plus the DC offset term to provide a baseband signal with real and imaginary portions.

A further object is to provide a device and method for VSB modulation of a digital signal in a DTV exciter which includes an interpolator for interpolating real and imaginary portions of a signal from a digital baseband filter so that a vestigial sideband modulated output signal includes third order intermodulation distortion products of the digital signal.

The present invention includes a method of vestigial sideband modulation for a digital signal, the method comprising the steps of:

(a) multiplying a digital signal with data symbols d_i input at rate $R = 1/T$, times $(\cos(\Pi i/2) - \sin(\Pi i/2))$;

EN:DOCID <EP 0853409A2 I >

(b) providing the product to a transversal filter with impulse response h_0 which provides separate sums of even and odd transversal filter coefficient terms;

(c) multiplexing the sums of even and odd coefficient terms to provide a baseband signal with real and imaginary portions $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$, in which

$$S_{bb}(kT) = \sum_i (d_i + b)\left(\cos(\frac{\Pi}{2}i) - j\sin(\frac{\Pi}{2}i)\right) h_0((k-i)T)$$

The invention also includes a filter for vestigial sideband modulation of a digital signal, comprising:

a multiplier for multiplying signal data symbols d_i input at rate $R = 1/T$, by $(\cos(\Pi i/2) - \sin(\Pi i/2))$;
a transversal filter with impulse response h_0 for receiving an output from said multiplier and for providing separate sums of even and odd transversal filter coefficient terms;
an adder for adding a DC offset term, $(\cos(\Pi i/2) - \sin(\Pi i/2))h_0(0)b$, to the sum of even coefficient terms; and
a multiplexer for multiplexing the sum of odd coefficient terms with the sum of even coefficient terms plus the DC offset term to provide a baseband signal with real and imaginary portions $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$, where

$$S_{bb}(kT) = \sum_i (d_i + b)\left(\cos(\frac{\Pi}{2}i) - j\sin(\frac{\Pi}{2}i)\right) h_0((k-i)T)$$

and, an interpolator for interpolating $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$ so that a vestigial sideband modulated output signal includes third order intermodulation distortion products of the signal.

The invention will now be described, by way of example, with reference to the accompanying drawings in which:

Figure 1 is a block diagram of an embodiment of the present invention.
Figure 2 is a graph depicting the frequency response $H_0(\omega)$ of a symmetrical lowpass shaping filter.
Figure 3 is block diagram of a further embodiment of the present invention.
Figure 4 is block diagram of a further embodiment of the present invention with ROMs instead of multipliers.
Figure 5 is a block diagram illustrating application of the present invention in a digital television exciter.
Figure 6 is partial block and partial circuit diagram showing complex signal flow in a filter for digital VSB modulation.
Figure 7 is a simplified block diagram of an embodiment of a baseband filter.
Figure 8 is a simplified block diagram of a further baseband filter.

Figure 1 depicts an embodiment of a vestigial sideband modulator that includes a baseband filter 10 which has a multiplier 12 for multiplying an input digital signal d_i arriving at rate $R = 1/T$ by $(\cos(\Pi i/2) - \sin(\Pi i/2))$, a transversal filter 14 with impulse response h_0 for receiving an output from multiplier 12 and for providing separate sums of even and odd filter coefficient terms, an adder 16 for adding a DC offset term to the sum of even coefficient terms, and a multiplexer 18 for multiplexing the sum of odd coefficient terms with the sum of even coefficient terms plus the DC offset term to provide a baseband signal with real and imaginary portions $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$, where

$$S_{bb}(kT) = \sum_i (d_i + b)\left(\cos(\frac{\Pi}{2}i) - j\sin(\frac{\Pi}{2}i)\right) h_0((k-i)T) \quad (1)$$

The preferred impulse response of transversal filter 14 is

$$h_0(t) = \Im(H_0(\omega)),$$

where $\Im$ is the Fourier Transform, and $H_0(\omega)$ is the frequency response of transversal filter 14 centered on zero frequency. The frequency response $H_0(\omega)$ of a symmetrical lowpass shaping filter, such as may be used for transversal filter 14 is seen in Figure 2.

The DC offset term added at adder 16 may be $(\cos(\Pi i/2) - \sin(\Pi i/2))h_0(0)b$, where b is a DC level added to the data

EPODOC ID: <EP _ 0853409A2_I_>

for the purpose of inserting some of the carrier frequency ω_0 into the VSB signal as a pilot carrier to support easy carrier recovery in a VSB receiver.

Figure 3 shows which numerical designations of features similar to those of Figure 1 have been retained. In transversal filter 14 pairs of data symbols which are to be weighted by a common coefficient have been combined in adders 20 before weighting by the common coefficient in multipliers 22. The registers 24 of transversal filter 14 have been "folded" about the central tap weight 26 around which the coefficient values are symmetrically disposed. This arrangement makes it easier to draw the symbol pairs which are to receive equal weight into their respective registers because of the transversal filter's even function impulse response.

Figure 4 depicts an alternative embodiment multipliers 22 replaced by read only memories (ROMs) 28. Multipliers may be replaced with ROMs as the inputs thereto are the sums of two n-bit (for 2^n-VSB) data symbols which can be defined by a (n + 1)-bit address which may be used to access from the ROM the result of multiplying the sum by the appropriate transversal filter coefficient. By using ROMs, the use of high speed multipliers or expensive DSP filter chips is avoided, thereby allowing high sampling rates. In an alternative embodiment, multiple addresses may be provided to a single ROM which provides the output previously provided by the sum of multiple ROMs.

The present invention finds application in an exciter for a digital television transmitter as illustrated in Figure 5. The baseband filter 30 is the embodiment of Figure 1, 3 or 4 for providing the real and imaginary signals 32. The frequency content of baseband filter 30 is out to a cutoff frequency of R/4 (R = 1/T) such as shown in Figure 2. The rate R imaginary signals represent the signal out to R/2 reasonably well. A further object of the invention is to represent the third order intermodulation products as well as the signal (for predistortion, and building digital linear and nonlinear correctors), three times the bandwidth of the signal is desirably represented. For the signal content of Figure 2, the signal must be represented out to a bandwidth of 3(R/4). To this end, an interpolator 34 for interpolating by two may be added to increase the sampling rate from R to 2R. Since the output rate is 2R, the signal is well represented out to a bandwidth of R, which is more than sufficient to represent the third order intermodulation products. The output from interpolator 34 may be provided to corrector 36 for linear and nonlinear correction which compensates for distortions downstream from the exciter. The signal from corrector 36 may then be provided to a further interpolator 38 which interpolates by two again and provides digital quadrature modulation onto an intermediate frequency (IF) carrier of frequency R. The output of interpolator 38 has a sampling rate of 4R. Interpolator 34 may be conventional (e.g., a Harris Corporation IC known as a Half Band Filter, HSP43216, operating in the interpolate-by-two mode), as may interpolator 38 which may be a similar half band filter which operates in a quadrature-to-real conversion mode.

The present invention provides numerous advantages, including an all digital implementation (no analog filters for sideband suppression), suitability for low-cost application specific integrated circuit (ASIC) implementation, use of standard ROMs in the embodiment of Figure 4 to allow high speed sampling, and support for high enough sampling rates to allow signal representation over three times the nominal signal bandwidth to thereby allow digital correction of third order intermodulation products. The present invention provides an output which is digitally correctable so that time-varying transmitter distortions can be tracked and adaptively corrected.

The following discussion starts with a complex signal representation for digital VSB modulation, as shown in Figure 6, and ends with the embodiment of Figure 1.

Figure 6 shows a DC offset b is added to incoming 8-level impulses. The DC offset is added to the data for the purpose of inserting some of the carrier frequency ω_0 into the VSB signal as a pilot carrier to support easy carrier recovery in the VSB receiver. The complex signal representation of the output VSB signal in Figure 6 is:

$$S_{vsb}(t) = \left[\left(\sum_i (d_i + b)\delta(t - iT)\right) \circledast \left(e^{j\left(\frac{2\pi}{4T}\right)t} h_0(t)\right)\right] e^{j\omega_0 t} \qquad (2)$$

where $\circledast$ represents convolution,

$$h_0(t) = \Im(H_0(\omega)),$$

$\Im$ is the Fourier Transform, $\delta(t)$ is the Dirac delta function or impulse function occurring at t = 0, and $H_0(\omega)$ is the frequency response of the shaping filter centered on zero frequency, such as shown in Figure 2. The factor $e^{j(2\pi/4T)t}$ accounts for the shaping filter's offset from the carrier by one-fourth the symbol rate of the modulator, and the factor

$$e^{j\omega_0 t}$$

represents the translation of the baseband spectrum to the vestigial sideband carrier frequency ω_0, which is also the

EN5DOCID <EP__ 0853409A2_I_>

199

~~198~~

pilot frequency.

As the exponential function is distributive across the convolution sign, equation 2 may be rewritten as:

$$S_{vsb}(t) = \left[\left(\sum_i (d_i + b)\delta(t-iT)e^{-j(\frac{2\Pi}{4T})t}\right) \circledast h_0(t)\right] e^{j(\omega_0 + \frac{2\Pi}{4T})t} \quad (3)$$

Equation 3 indicates that the DC offset input data signal (shown as the input in Figure 6) is to be downconverted by one-fourth the symbol rate by multiplying by $e^{j(2\Pi/4T)t}$, and passed through the lowpass filter of Figure 2 which has impulse response $h_0(t)$. The entire resultant baseband filtered signal (shown inside the brackets in equation 3) is upconverted and centered on a carrier frequency which is one-fourth symbol rate higher than the pilot frequency. The upconversion is represented by the exponential multiplier shown outside the brackets of equation 3. The convolution of equation 3 may then be simplified as shown in equation 4 below,
in which the last step is achieved by using the sifting property of the delta function which sifts out the value of the integrand at T = t-iT .

$$\left[\left(\sum_i (d_i+b)\delta(t-iT)e^{-j(\frac{2\Pi}{4T})t}\right) \circledast h_0(t)\right] = \int \left(\sum_i (d_i+b)\delta(t-\tau-iT)e^{-j(\frac{2\Pi}{4T})(t-\tau)}\right) h_0(\tau)d\tau \quad (4)$$
$$= \left(\sum_i (d_i+b)\right) e^{-j(\frac{2\Pi}{4T})iT} h_0(t-iT)$$

Substituting equation 4 into equation 3 and simplifying yields:

$$S_{vsb}(t) = \left[\sum_i (d_i+b)e^{-j(\frac{\Pi}{2})i} h_0(t-iT)\right] e^{j(\omega_0 + \frac{\Pi}{2T})t}$$
$$= \left[\sum_i (d_i+b)\left[\cos(\frac{\Pi}{2}i) - j\sin(\frac{\Pi}{2}i)\right] h_0(t-i\tau)\right] e^{j(\omega_0 + \frac{\Pi}{2T})t} \quad (5)$$

The baseband complex VSB signal is the portion of equation 5 within the outer brackets, which is the desired result (it is the same as equation 1.)

$$S_{bb}(kT) = \sum_i (d_i + b)\left(\cos(\frac{\Pi}{2}i) - j\sin(\frac{\Pi}{2}i)\right) h_0((k-i)T) \quad (6)$$

The signal of equations 1 and 6 may be provided by using a baseband filter such as shown in Figure 7. This filter may be simplified by noting that the real transversal filter 40 output depends only on odd numbered data symbols while the imaginary transversal filter 42 output depends only on even numbered data symbols. Further, it is noted that the real output alternates between a weighting of the even numbered coefficients of the transversal filter and a weighting of the odd numbered coefficients of the filter, and that the weighting of the imaginary output is opposite in sequence. Thus, the implementation of Figure 7 can be simplified to that of Figure 8.

The implementation of Figure 8 may be simplified to that of Figure 1 by changing the handling of the DC offset. The contribution of the DC offset to the sum of odd coefficients is zero for a filter with an odd number of taps, and that the contribution of DC offset to the sum of even coefficients is a sign-fluctuating constant which is proportional to the DC offset b and to the center tap of the even function transversal filter. The DC offset term added to the sum of even coefficient terms thus may be $(\cos(\Pi i/2)-\sin(\Pi i/2))h_0(0)b$.

A digital signal processing device and method for vestigial sideband (VSB) modulation of a digital signal. A binary digital signal d; is provided to a baseband shaping filter which samples the signal to provide a baseband signal which has real and imaginary portions $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$, where

EN5DOCID <EP 0853409A2 I >

$$S_{bb}(kT) = \sum_i (d_i + b)\left(\cos(\frac{\Pi}{2}i) - j\sin(\frac{\Pi}{2}i)\right)h_0((k-i)T)$$

The VSB signal is provided with perfect sideband suppression without using analog filters.

Claims

1. A method of vestigial sideband modulation for a digital signal, the method comprising the steps of:

   (a) multiplying a digital signal with data symbols d_i input at rate $R = 1/T$, times $(\cos(\Pi i/2)-\sin(\Pi i/2))$;
   (b) providing the product to a transversal filter with impulse response h_0 which provides separate sums of even and odd transversal filter coefficient terms;
   (c) multiplexing the sums of even and odd coefficient terms to provide a baseband signal with real and imaginary portions $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$, in which

$$S_{bb}(kT) = \sum_i (d_i + b)\left(\cos(\frac{\Pi}{2}i) - j\sin(\frac{\Pi}{2}i)\right)h_0((k-i)T)$$

2. A method as claimed in Claim 1 including the step of interpolating $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$ to provide a sampling rate of at least 2R so that a vestigial sideband modulated output signal includes third order intermodulation distortion products of the digital signal, and the step of adding a DC offset term, $(\cos(\Pi i/2)-\sin(\Pi i/2))h_0(0)b$, to the sum of even coefficient terms, before the multiplexing step.

3. A method as claimed in Claim 1 or 2 including the steps, in the transversal filter, of combining pairs of input data symbols which are to be weighted with a common filter coefficient, and thereafter weighting the combined pairs with their common filter coefficient, in which the step of weighting the combined pairs comprises the step of addressing read only memories which contain addressable products of filter coefficients times input data symbol values.

4. A method as claimed in Claim 3 including the step of providing an array of registers for the input data symbols, the registers extending in rows from a central tap so that registers for pairs which are to be combined are adjacent.

5. A method as claimed in any one of Claims 1 to 4 wherein the digital signal is a television signal.

6. A method of vestigial sideband modulation for a digital signal, the method comprising the steps of:

   (a) providing an input digital signal at a rate 1/T;
   (b) downconverting the input digital signal by multiplying it by $e^{j(2\Pi/4T)t}$;
   (c) providing the downconverted digital signal to a symmetrical shaping filter which Nyquist filters the downconverted digital signal and has an impulse response of $h_0(t)$; and
   (d) upconverting the filtered digital signal so that it is centered on a carrier frequency, ω_0, in the vestigial sideband of the input digital signal by multiplying the filtered digital signal by

$$e^{j(\omega_0 + (2\Pi/4T))t}.$$

7. A method as claimed in Claim 6 including the step of separating the upconverted signal with real and imaginary portions $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$, wherein

$$S_{bb}(kT) = \sum_i (d_i + b)\left(\cos(\frac{\Pi}{2}i) - j\sin(\frac{\Pi}{2}i)\right)h_0((k-i)T)$$

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and, the step of interpolating $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$ to provide a vestigial sideband modulated output signal which includes third order intermodulation distortion products of the digital signal, with the filter providing separate sums of even and odd coefficient terms.

8. A method as claimed in Claims 6 or 7 wherein the steps of adding a DC offset term to the sum of even coefficient terms, and multiplexing the sum of odd coefficient terms with the sum of even coefficient terms plus the DC offset term to provide $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$, and in the transversal filter, of addressing read only memories which contain addressable products of filter coefficients times input values, in which the digital signal is a television signal.

9. A filter for vestigial sideband modulation of a digital signal, comprising:

a multiplier for multiplying signal data symbols d_i input at rate $R = 1/T$, by $(\cos(\Pi i/2)-\sin(\Pi i/2))$;
a transversal filter with impulse response h_0 for receiving an output from said multiplier and for providing separate sums of even and odd transversal filter coefficient terms;
an adder for adding a DC offset term, $(\cos(\Pi i/2)-\sin(\Pi i/2))h_0(0)b$, to the sum of even coefficient terms; and
a multiplexer for multiplexing the sum of odd coefficient terms with the sum of even coefficient terms plus the DC offset term to provide a baseband signal with real and imaginary portions $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$.
where

$$S_{bb}(kT) = \sum_i (d_i + b)\left(\cos(\frac{\Pi}{2}i) - j\sin(\frac{\Pi}{2}i)\right)h_0((k-i)T)$$

and, an interpolator for interpolating $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$ so that a vestigial sideband modulated output signal includes third order intermodulation distortion products of the signal.

10. A filter as claimed in Claim 9 wherein said transversal filter comprises plural adders for combining pairs of input data symbols which are to be weighted with a common filter coefficient, in which said transversal filter comprises read only memories which contain addressable products of filter coefficients times an input values.

11. A filter as claimed in Claim 10 wherein said transversal filter comprises an array of plural registers for providing input data symbols to said adders, said registers extending symmetrically from a central tap so that pairs of ones of said registers which hold input data symbols which are to be combined are adjacent.

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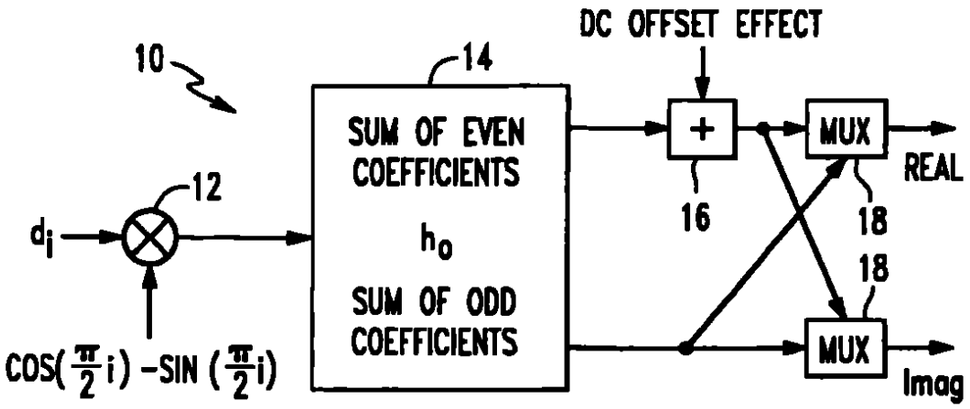


FIG. 1

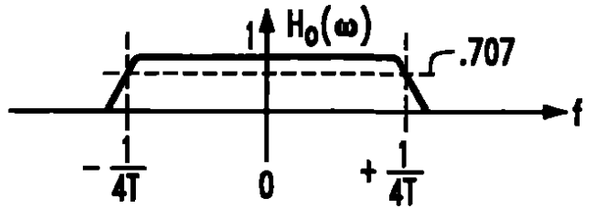


FIG. 2

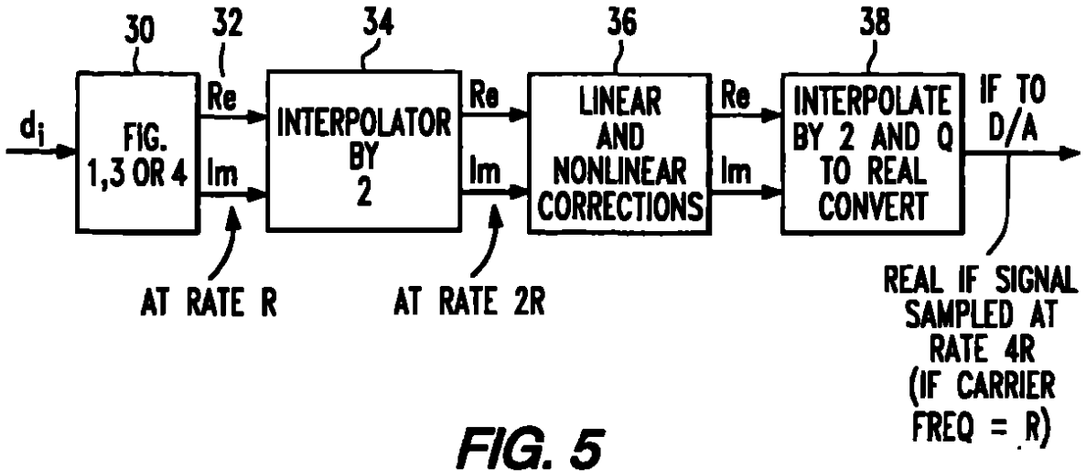


FIG. 5

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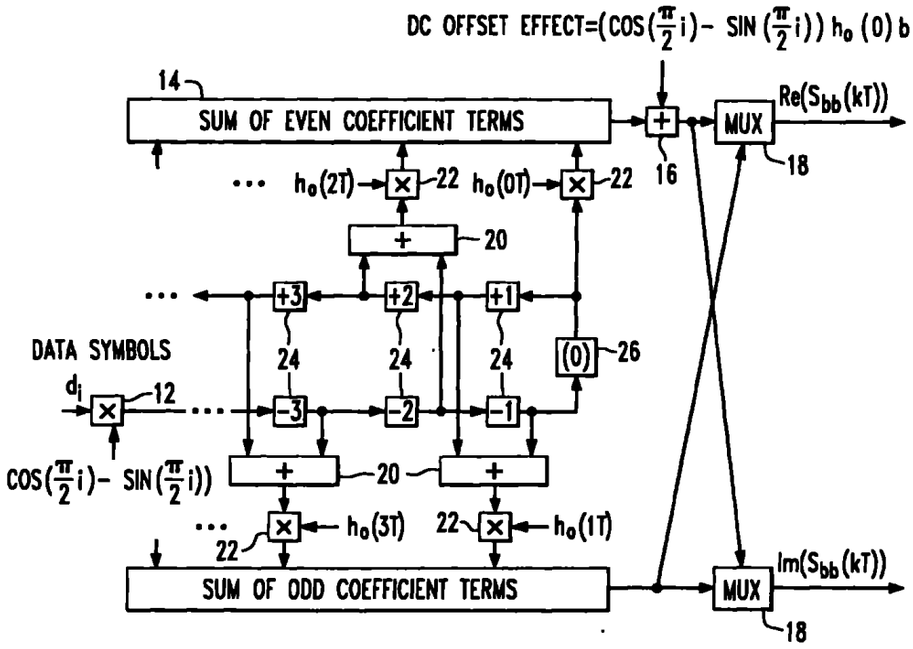

DC OFFSET EFFECT=(COS(π/2 i)- SIN (π/2 i)) h₀ (0) b
14
SUM OF EVEN COEFFICIENT TERMS
Re(S_bb(kT))
MUX
18
16
h₀(2T)
22
h₀(0T)
22
20
+3
+2
+1
24
24
24
(0)
26
DATA SYMBOLS
d_i
12
-3
-2
-1
COS(π/2 i)- SIN(π/2 i))
20
20
22
h₀(3T)
22
h₀(1T)
SUM OF ODD COEFFICIENT TERMS
Im(S_bb(kT))
MUX
18

FIG. 3

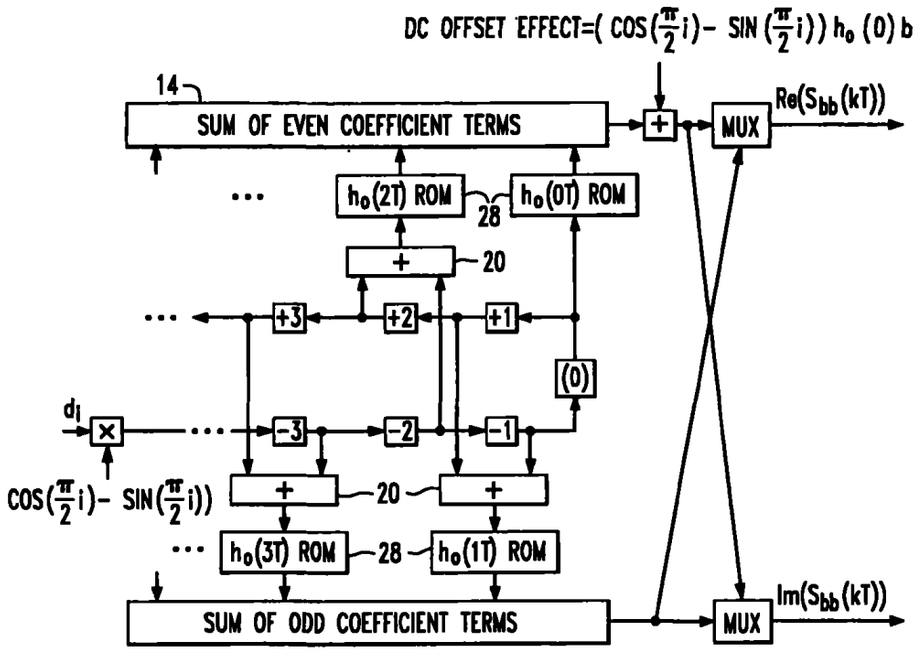

DC OFFSET EFFECT=(COS(π/2 i)- SIN (π/2 i)) h0 (0) b
14
SUM OF EVEN COEFFICIENT TERMS
+
MUX
Re(Sbb(kT))
h0(2T) ROM
28
h0(0T) ROM
20
+3
+2
+1
(0)
di
-3
-2
-1
COS(π/2 i)- SIN(π/2 i))
h0(3T) ROM
h0(1T) ROM
SUM OF ODD COEFFICIENT TERMS
Im(Sbb(kT))

FIG. 4

202

201

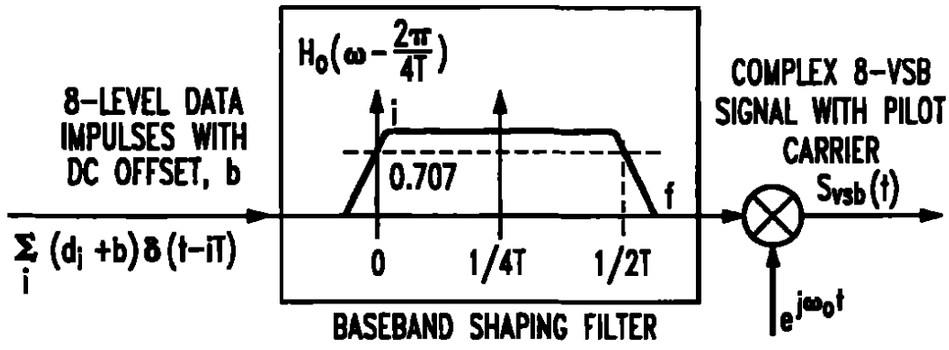


FIG. 6

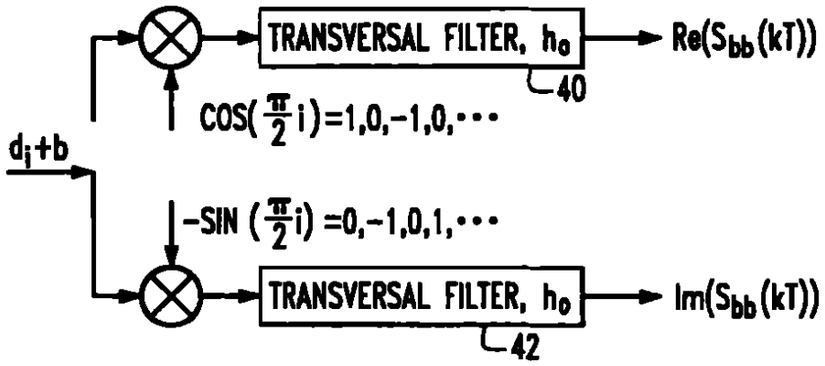


FIG. 7

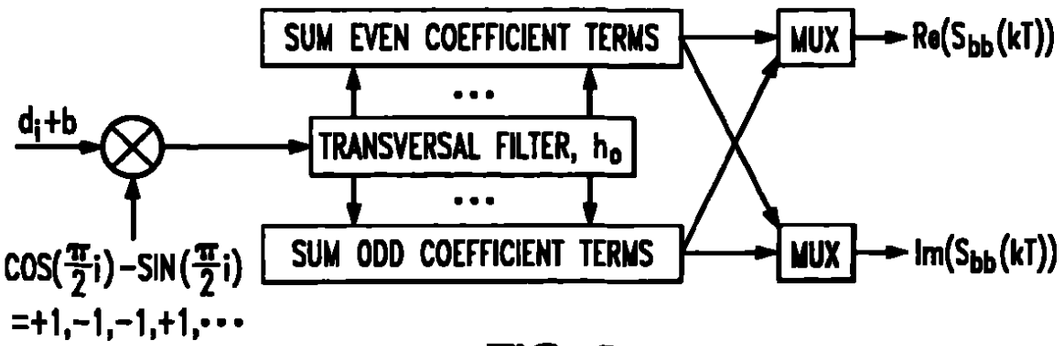


FIG. 8

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(54) **Device and method for digital vestigial sideband modulation**

(57) A digital signal processing device and method for vestigial sideband (VSB) modulation of a digital signal. A binary digital signal d_l is provided to a baseband shaping filter which samples the signal to provide a baseband signal which has real and imaginary portions $Re(S_{bb}(kT))$ and $Im(S_{bb}(kT))$, where

$$S_{bb}(kT) = \sum_l (d_l + b)\left(\cos(\frac{\Pi}{2}l) - j\sin(\frac{\Pi}{2}l)\right)h_0((k-l)T)$$

The VSB signal is provided with perfect sideband suppression without using analog filters.

EP 0 853 409 A3

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 97 31 0657

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The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
BERLIN	8 December 2000	Farese, L

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
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A : technological background
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P : intermediate document

T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

EPO FORM 1503 03.82 (P04C01)

EN5DOCID <EP 0853409A3 I >

204
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ANNEX TO THE EUROPEAN SEARCH REPORT ON EUROPEAN PATENT APPLICATION NO. EP 97 31 0657

This annex lists the patent family members relating to the patent documents cited in the above-mentioned European search report. The members are as contained in the European Patent Office EDP file on
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ANNEX TO THE EUROPEAN SEARCH REPORT ON EUROPEAN PATENT APPLICATION NO. EP 97 31 0657

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United States Patent [19]

Nossek

[11] **Patent Number: 4,617,537**

[45] **Date of Patent: Oct. 14, 1986**

[54] **METHOD FOR DIGITAL QUADRATURE AMPLITUDE MODULATION**

[75] Inventor: **Josef A. Nossek, Iffeldorf, Fed. Rep. of Germany**

[73] Assignee: **Siemens Aktiengesellschaft, Berlin and Munich, Fed. Rep. of Germany**

[21] Appl. No.: **595,527**

[22] Filed: **Mar. 30, 1984**

[30] **Foreign Application Priority Data**

Apr. 22, 1983 [DE] Fed. Rep. of Germany 3314603

[51] Int. Cl.⁴ H03C 1/00; H03D 1/00

[52] U.S. Cl. ..... 332/48; 332/9 R; 332/31 R

[58] Field of Search ..... 332/9 R, 31 R, 41, 42, 332/48; 375/41, 43; 455/108, 109

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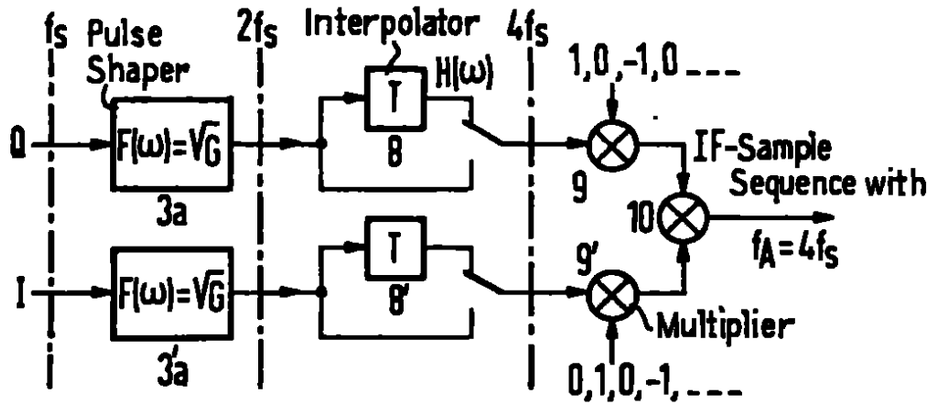
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Primary Examiner—Eugene R. LaRoche
Assistant Examiner—Robert J. Pascal
Attorney, Agent, or Firm—Hill, Van Santen, Steadman & Simpson

[57] **ABSTRACT**

Spectrum-converting digital filters are employed for quadrature amplitude modulation. Given the method of the invention, the frequency of the two carrier oscillations residing at 90° relative to one another amounts to one-fourth the sampling frequency and, in addition, their phase positions relative to the sampling times is selected such that every second sample of both carrier oscillations is alternately zero.

1 Claim, 6 Drawing Figures



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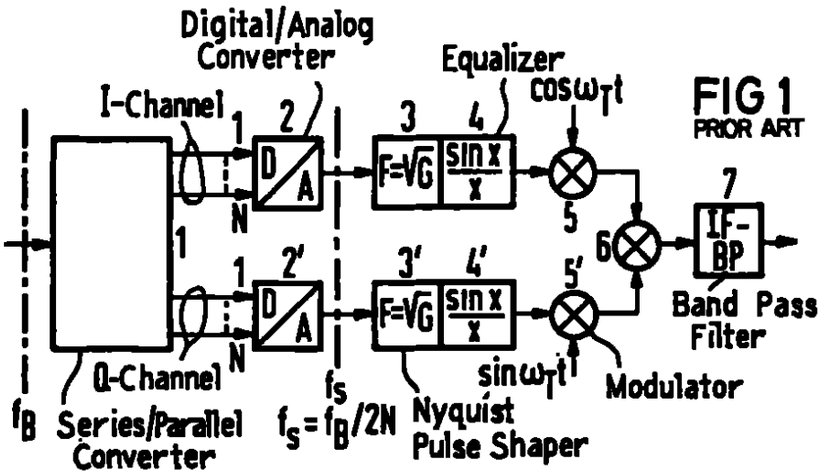


FIG 1
PRIOR ART

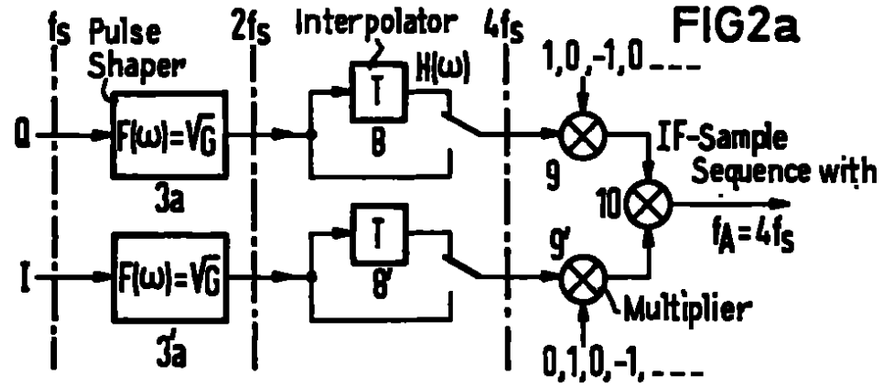


FIG 2a

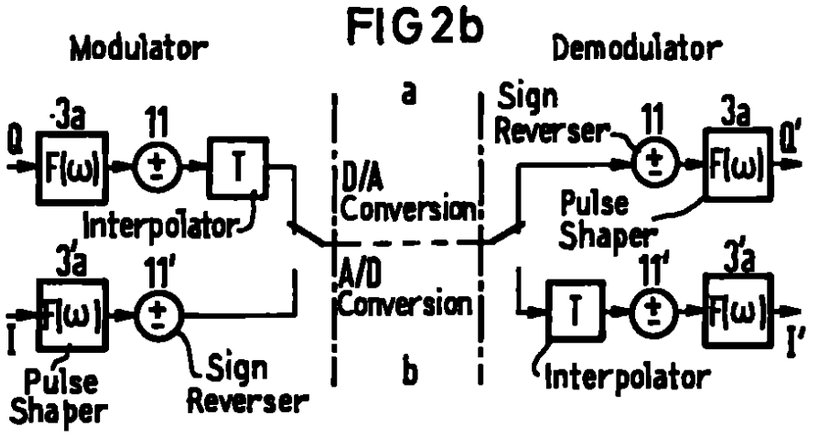


FIG 2b

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FIG 3a

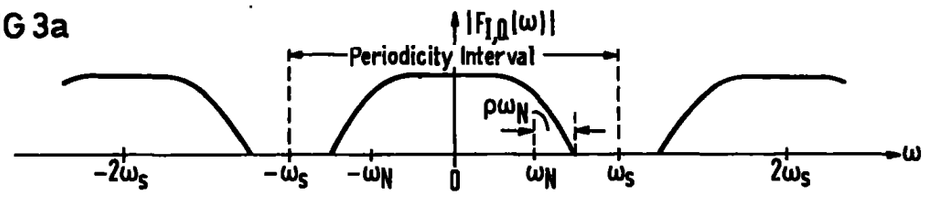


FIG 3b

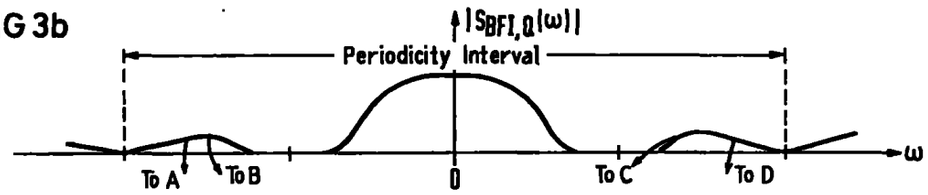
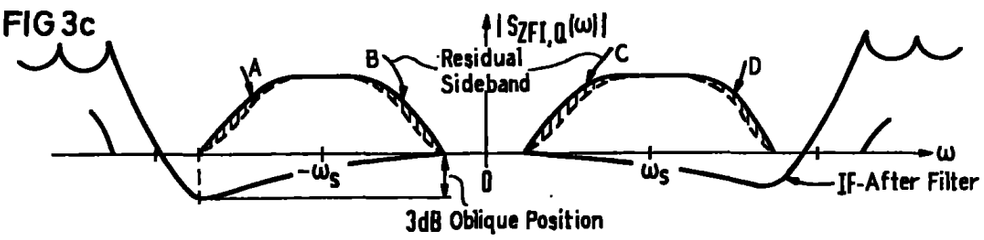


FIG 3c



METHOD FOR DIGITAL QUADRATURE AMPLITUDE MODULATION

BACKGROUND OF THE INVENTION

1. Field of the Invention

The present invention relates to a method of digital quadrature amplitude modulation using spectrum-shaping digital filters wherein sampling sequences consisting of two carrier oscillations residing at 90° relative to one another are available.

2. Description of the Prior Art

In terms of basic concept, methods for quadrature amplitude modulation (QAM) have been disclosed in the publication NTG Fachberichte (1980), pp. 81–85. Digital, spectrum-covering low-pass filters are also employed in this reference. The digital/analog conversion subsequently occurs and the quadrature amplitude modulation occurs on an analog basis. This also applies, analogously, to the receiving path.

It should be further taken into consideration that circuits that meet the so-called Nyquist condition are required for the resolution of a number of telecommunications transmission problems. This condition is defined per se in that appertaining circuits are required when, for example, it is a matter of transmitting pulses, without intersymbol noise (i.e. pulses without crosstalk) or, respectively, of converting video frequency bands, whereby the frequency zero and a vestigial sideband must also be transmitted. Possibilities for the synthesis of Nyquist filters are specified for processing analog signals, for example, in the periodical "Frequenz", 1973, Vol. 27, pp. 2–6. In terms of basic concept, these circuits amount to the fact that realizing so-called strict diplexers wherein filters having a so-called self-reciprocal characteristic function are employed. In the meantime, so-called digital technology is also employed in telecommunications transmission systems and, therefore, the problem also arises of creating selection devices therefore, i.e. circuits and networks for which the term digital filters has become standard. Possibilities for the realization of such digital filters, for example, are set forth in rather general terms in the book by Temes and Mitra, entitled "Modern Filter Theory and Design", John Wiley & Sons, 1973, particularly on Pages 505–557. So-called wave digital filters comprise a special species of these filters. Such circuits are disclosed, for example, in the German Letters Pat. No. 2,027,303 or, respectively, in the periodical "AEÜ", Vol. 25 (1971), pp. 79–89. Wave digital filters have the advantage that analog reference filters can be directly digitally simulated.

As already mentioned above, the quadrature amplitude modulation in the initially-mentioned article (NTG Fachberichte, pp. 81–88) is carried out on an analog basis.

SUMMARY OF THE INVENTION

The object of the present invention is to provide specific possibilities with which all further method steps can be digitally executed upon the use of spectrum-converting digital filters.

Given methods of the type mentioned above, this object is achieved, according to the present invention, in such a system which is characterized in that the frequency of two carrier oscillations which are at 90° relative to one another is one-fourth of the sampling frequency and their phase positions relative to the sampling times is selected such that every second sample of both carrier oscillations is alternately zero in such a manner that respectively one of the two carrier oscillations is zero at every clock time.

BRIEF DESCRIPTION OF THE DRAWINGS

Other objects, features and advantages of the invention, its organization, construction and operation will be best understood from the following detailed description, taken in conjunction with the accompanying drawings, on which:

FIG. 1 is a schematic representation of a known system which is provided for a better understanding of the present invention;

FIG. 2*a* is a schematic representation of a QAM modulator, having a pulse shaper, for carrying out the present invention;

FIG. 2*b* is a schematic representation of a QAM modulator and demodulator in which an interpolator or, respectively, a decimator are combined;

FIG. 3 is a graphic illustration of wave forms with respect to frequency ω, wherein further;

FIG. 3*a* illustrates the baseband spectrum following the pulse shaper;

FIG. 3*b* illustrates the baseband spectrum following an interpolation filter; and

FIG. 3*c* illustrates the intermediate frequency spectrum (IF spectrum) following a multiplier and the fundamental attenuation curve of an IF after-filter for the correction of the sinx/x distortion of the digital/analog converter and for the suppression of the periodic continuation of the spectrum.

DESCRIPTION OF THE PREFERRED EMBODIMENTS

Referring to FIG. 1, a bit frequency f_B is input into a serial/parallel converter 1. A pair of channels I and Q are generated in which respective stages 1–N are indicated. The digital/analog converters 2 and 2', respectively, follow in the channel I and in the channel Q and the so-called symbol frequency f_S can be subsequently seen, whereby the relationship $f_S=f_B/2N$ holds true and 2_N-level signals are present in the channel I and in the channel Q. Identical circuits for the Nyquist pulse shaping follow, these being provided with reference characters 3 and 3', respectively, and having the transfer function $F=\sqrt{G}$. The sinx/x equalizers 4 and 4' follow so that hold distortions of the digital/analog converter can be intercepted. The Nyquist pulse shaping occurs at 50% of the pulse in the modulator and these modulators are referenced 5 in the channel I and 5' in the channel Q and the 90° phase shift of these two channels is achieved by way of multiplication factors cos $\omega_T \cdot T$ or, respecby way of multiplication factors cos $\omega_T \cdot T$ or, respectively, sin $\omega_T \cdot T$. The quadrature modulators 5 and 5' are therefore multipliers and the channels I and Q are recombined in an adder 6 and supplied to a following intermediate frequency bandpass filter 7. This IF bandpass filter then filters out undesired modulation products. Given a resolution of N bits, the digital/analog converters 2 and 2', respectively, must have a considerably-higher accuracy, in particular M=N+(4...6) bits, whereby M signifies the accuracy. The conversion rate is thereby equal to the sample frequency f_S.

Given the traditional quadrature modulator having pulse shaping according to FIG. 1, the serial bit stream is first converted into two multi-level symbol streams in

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a serial/parallel converter 1 and is subsequently digital-/analog converted (2,2'). Given a 16 QAM method, the resolution of these digital/analog converters amounts, for example, to 2 bits in each quadrature channel. The accuracy of these converters, however, must be considerably higher than the resolution (estimated at about 4 . . . 6 bits, i.e. higher by the factor 16 . . . 64). The following generally applies.

A 2^{2N} QAM method is based on 2^N-level signals in the quadrature channels I,2 and on an N-bit resolution and an M=(N+(4 . . . 6)) bit accuracy of the digital-/analog converter. The pulse shaping at the transmitting side subsequently occurs, this together with the pulse shaping to be identically provided at the receiving side (demodulator), meeting the Nyquist condition type 1 and the intersymbol noise thus disappearing. This identical division to modulator and demodulator is optimum in view of the adjacent-channel influence. Furthermore, a sin x/x equalization 4, 4' is co-incorporated into the pulse shaper at the transmitting side in order to equalize the hold distortions of the digital/analog converters.

The following requirement is made of the carrier frequency ω at the following modulation (multiplication 5,5'):

$$\omega_T=2\pi f_T\geq(1+\zeta)2\pi f_N;\ f_N=2f_N \quad (1)$$

where ζ is the known roll-off factor and f_n is the Nyquist frequency of the pulse shaper.

In the transition to a digital concept, it is necessary for reasons of realizability that the intermediate frequency (IF) carrier be selected as low as possible upon consideration of equation (1).

The IF selection (bandpass filter 7) occurs after the combination of the two quadrature channels I and Q in order to suppress undesired, higher modulation products and interference lines.

The demodulator proceeds directly from the modulator by way of reversing the direction of signal flow. The sin x/x equalizer in the pulse shaper is then omitted. Added thereto is the carrier reacquisition and the clock reacquisition that are of importance in the coherent demodulation and in the regeneration of the characters. No fixed relationship between carrier and clock is usually provided given the conventional principle of construction.

Here, also, there is a significant difference in the transition of the digital design. The sampling frequency at the input of the modulator (multiplier) must be a whole multiple of both the character frequency f_S and of the carrier frequency f_T.

$$f_A=K_1f_S=K_2f_T\in K_1,\ K_2\in N. \quad (2)$$

Only the case of bandband pulse shaping has been considered, since it leads to the lowest operating speed given a digital solution.

FIGS. *2a* and *2b* include circuit elements that have already been explained with reference to FIG. 1. The channels Q and I may again be seen, along with a so-called interpolating pulse shaper 3*a* or 3'*a*, which is followed in FIG. *2a* by an interpolator having the transfer function $H(\omega)$. These interpolators are indicated with the reference characters 8 and 8' and are illustrated as delay elements T followed by switches. A multiplier 9 is connected to the interpolator 8 in the channel Q and a multiplier 9' is connected by way of a respective switch to the interpolator 8' in the channel I. The interpolator 9 multiplies the signal sequence by the numerical sequence 1, 0, −1, 0, . . . when the multiplier 9' in the channel I multiplies the signal sequence by the numerical sequence 0, 1, 0, −1.

The symbol frequency f_S or, respectively, $2f_S$ or, respectively, $4f_S$ is also identified in FIG. *2a* with broken lines, i.e. at those locations at which the interpolation takes effect.

Interpolator and decimator are combined in FIG. *2b* and the arrangement is indicated for the modulator and demodulator sides. As likewise symbolically illustrated, the reference characters 11 and 11', respectively, indicate digital circuits that carry a reversal of operational sign. A likewise indicated by broken lines, the digital-/analog and, respectively, analog/digital conversions and other conversions can be undertaken between the modulator and the demodulator.

The baseband spectrum after the pulse shaper is shown in FIG. *3a* as a function of the frequency ω, as is the bandband spectrum after the interpolation filter in FIG. *3b*.

The periodicity intervals are likewise indicated and further designations shall be explained below. The IF spectrum after the multiplier is illustrated in FIG. *3c*, as is a basic attenuation curve of an IF subsequent filter for the correction of the sin x/x distortion of the digital-/analog converter and for the suppression of the periodic continuation of the spectrum.

It is advantageous when the spectrum-shaping filters preceding the modulators 9, 9' are interpolating or, respectively, decimating bridge wave digital filters fulfilling the Nyquist condition that simultaneously execute the reversal of the operational sign of every second sample. Such bridge wave digital filters are disclosed in a copending application Ser. No. 595,400, filed Mar. 30, 1984.

For the purpose of a further, general explanation of FIGS. *2a* and *2b* or, respectively, FIG. 3, the following should also be noted in comparison to FIG. 1.

Basic significance is accorded to the selection of the constants K_1 and K_2 according to equation (2) upon consideration of equation (1). In addition, it is advantageous when the samples of the sinusoidal carry oscillation only assume the values +1, 0and −1. This is the case when the constant K_2 assumes the values

$$K_2=2,\ 3,\ 4 \text{ or } 6.$$

Of these four possibilities, $K_2=4$ is an excellent value, since the corresponding sampling sequences for the two carrier oscillations

$$\cos \omega_T nT\text{: } 1,0,-1,0,1,0,-1,\ldots$$

$$\sin \omega_T nT\text{: } 0,1,0,-1,0,1,0,\ldots$$

with $T=1/f_A$, $f_A=4f_T$

and exhibits only a sampling value differing from zero (+or−) at every sampling time and the combination of the two quadrature channels I and Q therefore becomes particularly simple. This means that the quadrature modulation is accomplished by way of simple, alternating blanking of every second sample and through-connection or, respectively, inversion of the remaining values in the quadrature channels, and the combination is accomplished by way of interleaving the two sampling sequences.

In order to observe equation (1) for all $\zeta\in[0,1]$ it suffices to select

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$f_S = f_T = 1/4\ f_A$

i.e. $K_1 = K_2 = 4$ and equation (2) is likewise fulfilled.

For example, sampling sequences having $f_A = 4f_S$ must be supplied to the modulators (multipliers). This means that the pulse-shaping filters must interpolate, i.e. must increase the sampling rate by the factor 4. This interpolation should advantageously be executed in two stages by, respectively, the factor 2. In the following, the spectrum or, respectively, pulse shaping is thereby completely shifted into the first part of the sampling rate increase (or, respectively, reduction, i.e. decimation in the demodulator).

The second part of the sampling rate change therefore relates to a sequence whose spectrum has already experienced a 50% Nyquist shaping and should accomplish the sampling rate 4 f_S required for the modulation with the least expense possible. As shown below, this can be executed in a particularly efficient fashion with a type of vestigial sideband modulation for both quadrature channels.

According to FIG. 2a, an interpolation filter 8, 8' having the transfer function

$$H(\omega) = 1 + e^{-j\omega T} = 2e^{-j\omega\frac{T}{2}} \cos\omega\frac{T}{2} \tag{3}$$

is employed. Then, deriving for the baseband spectrum at the multiplier input in the channel I, as a consequence of a symbol is

$$S_{BF}(\omega) = 2e^{-j\omega\frac{T}{2}} \cos\omega\frac{T}{2} \cdot F(\omega) \tag{4}$$

The corresponding case applies to the channel Q. Deriving for the IF spectra before the combination is

$$S_{ZF}(\omega) = \frac{1}{2}\left(S_{BF}(\omega - \omega_2) + S_{BF}(\omega + \omega_2)\right) \tag{5}$$

$$S_{ZFQ}(\omega) = \frac{1}{2j}\left(S_{BFQ}(\omega - \omega_2) - S_{BFQ}(\omega + \omega_2)\right).$$

Using the periodicity of $F_{I,Q}(\omega)$, equation (4) is inserted in equation (5) and produces

$$F_{I,Q}(\omega) = F_{I,Q}(\omega + 2n\omega_2),\ n \in Z. \tag{6}$$

For the IF spectra

$$S_{ZFI} = F(\omega - \omega_2) \tag{7}$$

$$S_{ZFQ} = je^{-j(\omega-\omega_2)T} F_Q(\omega - \omega_2).$$

and the following applies

$$S_{ZF}(\omega) = S_{ZFI}(\omega) + S_{ZFQ}(\omega) = F(\omega - \omega_2) + je^{-j(\omega-\omega_2)T} F_Q(\omega - \omega_2). \tag{8}$$

The result shows that the IF spectrum arises by way of simple shifting of the spectrum at the pulse shaped output (and potential phase rotation) and that the interpolation filter causes no distortions, even though it has a cosine-shaped frequency response. This is to be attributed to a type of vestigial sideband modulation as illustrated in FIG. 3 with reference to the spectra.

FIG. 3c additionally shows the fundamental frequency response of a time-continuous analog IF filter that equalizes the sin x/x distortions of the digital-/analog converter in its pass band and suppresses the periodic continuation of the spectrum in its stop band. The sin x/x compensation could potentially be undertaken in the digital part as well.

The derivation shown here for the modulator applies by full analogy to the demodulator, including the decimator.

Although I have described my invention by reference to particular illustrative embodiments thereof, many changes and modifications of the invention may become apparent to those skilled in the art without departing from the spirit and scope of the invention. I therefore intend to include within the patent warranted hereon all such changes and modifications as may reasonably and properly be included within the scope of my contribution to the art.

I claim:

1. A method for clocked digital amplitude modulation, wherein signal sampling sequences are available from two carrier oscillations which are at 90° phase positions relative to one another and at a predetermined sampling frequency, comprising the steps of:

producing the carrier oscillations at one-fourth of the predetermined sampling frequency;

setting the phase positions relative to the sampling times such that every second sample of both carrier oscillations is alternately zero so that respectively one of the two carrier oscillations is zero at every clock time;

reversing the operational sign of every second sample to provide two sampling sequences at half the sampling frequency; and

interleaving the two sampling sequences to provide quadrature amplitude modulation offset by one sampling.

* * * * *

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US005696796A

United States Patent [19]

Poklemba

[11] Patent Number: 5,696,796

[45] Date of Patent: Dec. 9, 1997

[54] **CONTINUOUSLY VARIABLE IF SAMPLING METHOD FOR DIGITAL DATA TRANSMISSION**

[75] Inventor: **John James Poklemba**, Ijamsville, Md.

[73] Assignee: **Comsat Corporation**, Bethesda, Md.

[21] Appl. No.: **483,764**

[22] Filed: **Jun. 7, 1995**

[51] Int. Cl.[6] **H03K 9/00**; H04L 27/06; H04L 27/14; H04L 27/22

[52] U.S. Cl. **375/316**; 332/103

[58] Field of Search 375/316, 326; 332/103; 329/129

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Primary Examiner—Stephen Chin

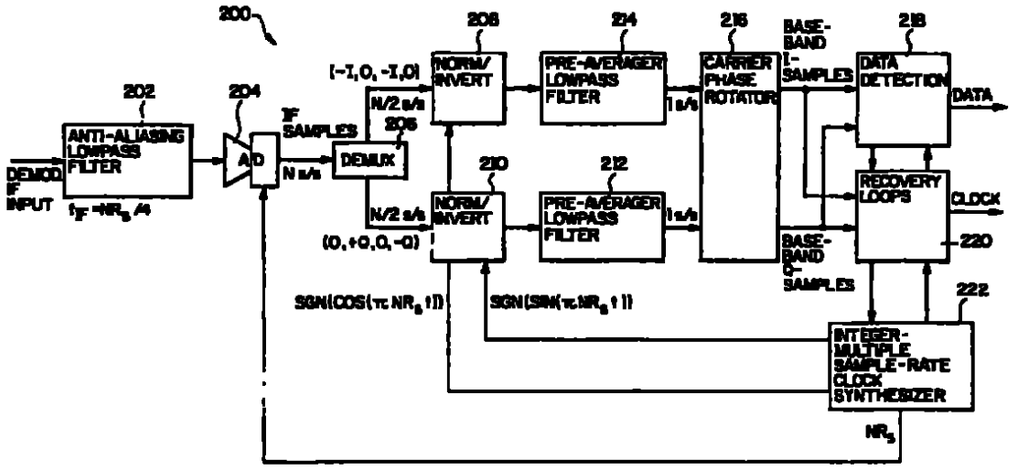
Assistant Examiner—Mohammad Ghayour

Attorney, Agent, or Firm—Sughrue, Mion, Zinn, Macpeak & Seas, PLLC

[57] **ABSTRACT**

A sampling method, denoted continuously variable IF sampling (CVIFS), implemented in digital transmission and reception components is characterized by sampling a bandpass signal which is centered at an intermediate frequency f_{IF} that is related to both the data transmission symbol rate (R_s) and sample rate (NR_s) by integer ratios. Moreover, the CVIFS method is characterized in that the sample rate NR_s is constrained to be four times the IF center frequency, and both the sample rate NR_s and the IF center frequency are constrained to be within a range of a factor of 1.5 of their maximum values. To maintain this high sample rate as the data symbol rate is lowered, a "gear shifting" technique is utilized whereby the integer number of samples-per-symbol, N, is increased as the symbol rate decreases. Moreover, the sample rate NR_s is always between 3.0 to 4.5 times the maximum data symbol rate (R_{max}), and the IF center frequency is always between 3/4 to 9/8 times (R_{max}), irrespective of the actual data rate R_s.

6 Claims, 8 Drawing Sheets



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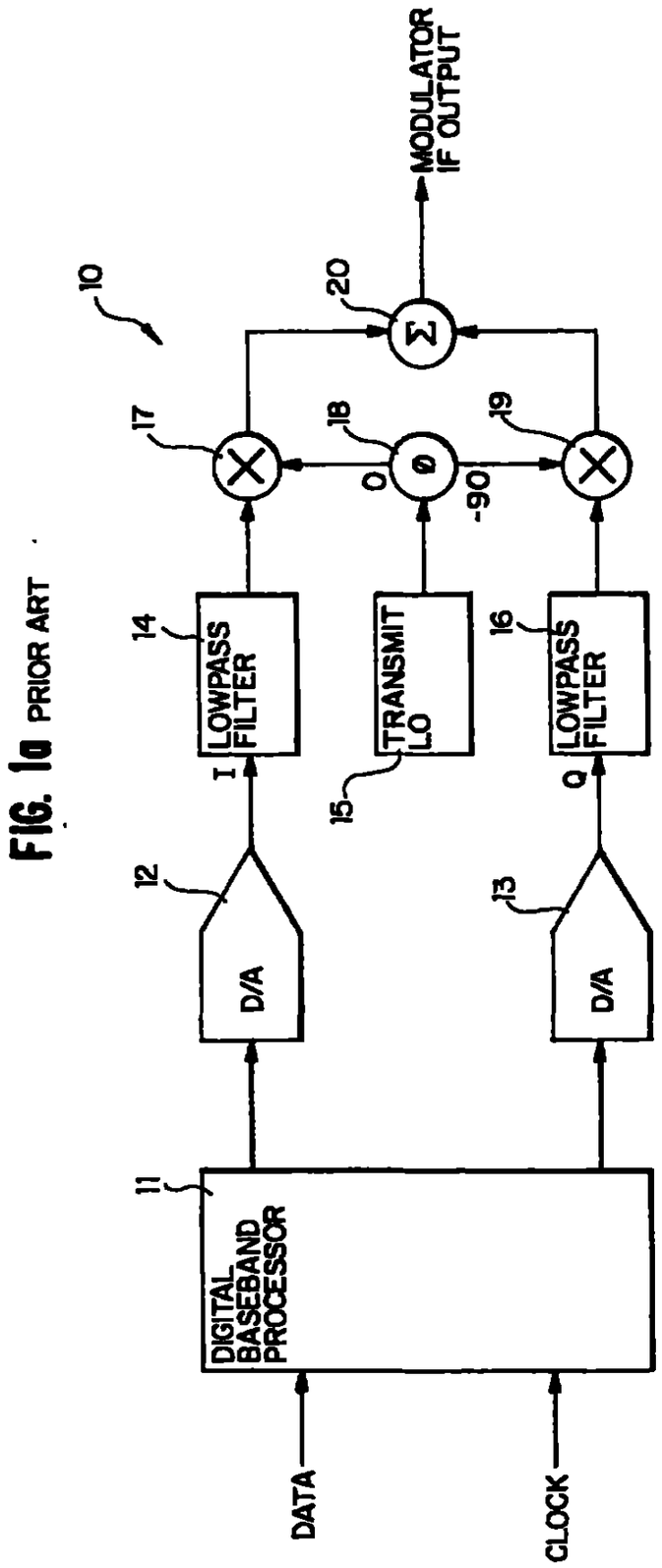


FIG. 1a PRIOR ART

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FIG. 1b PRIOR ART

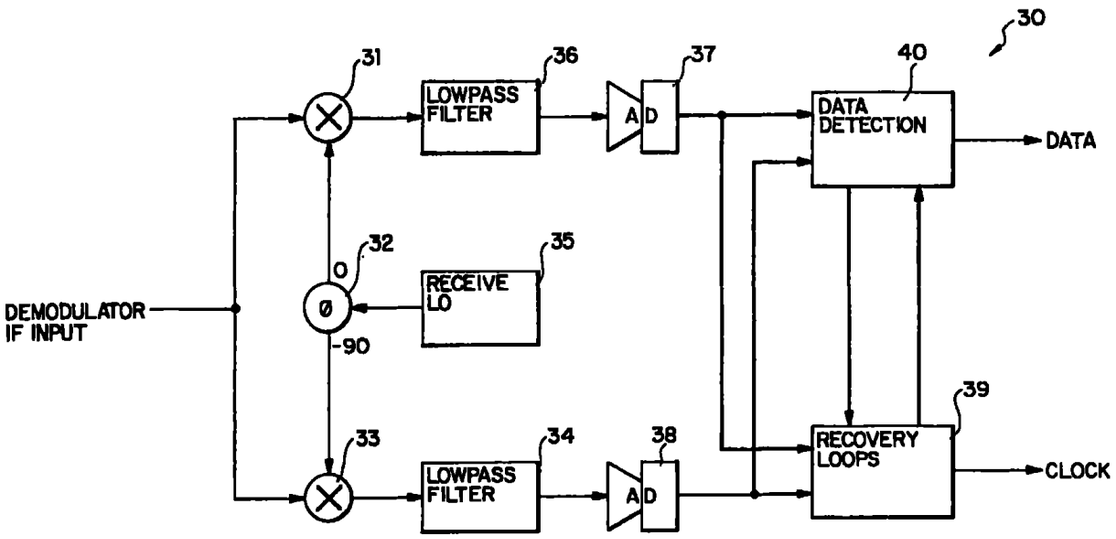
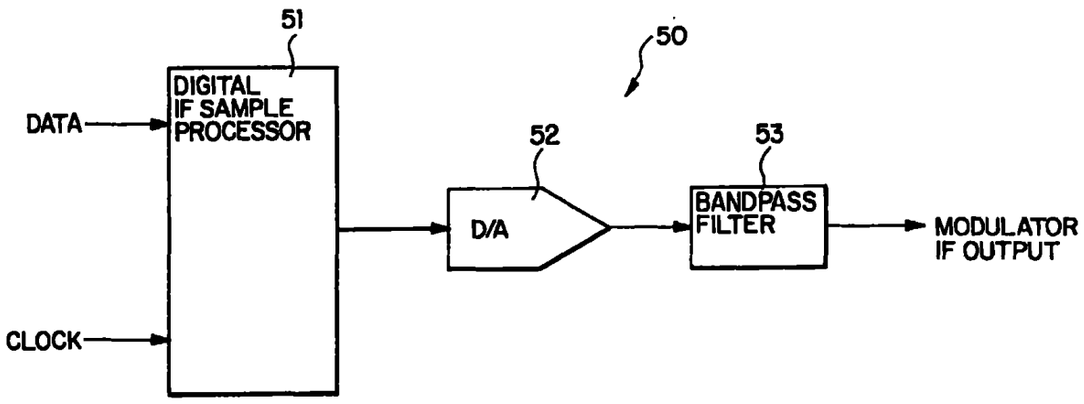
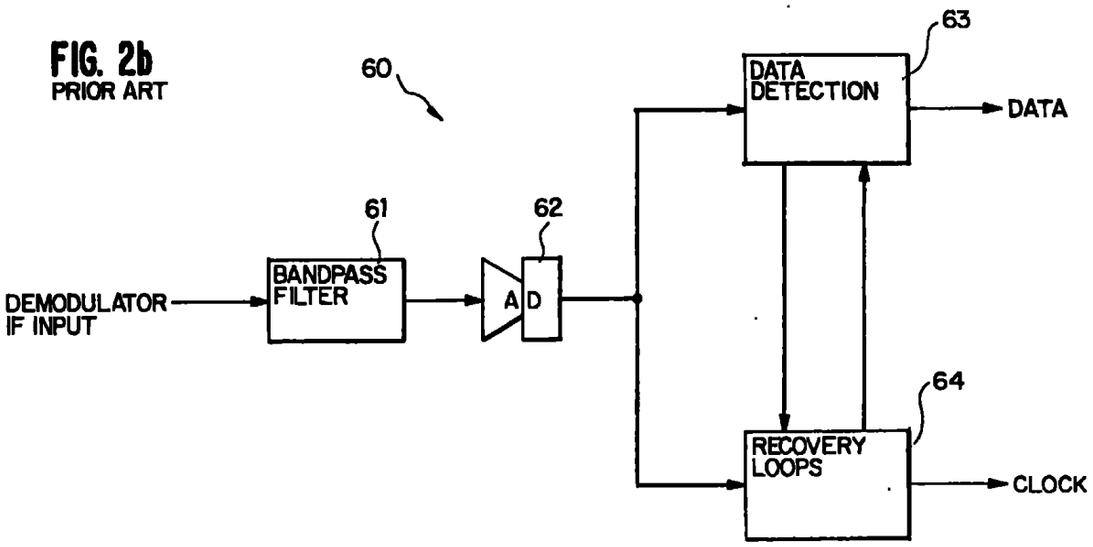


FIG. 2a PRIOR ART

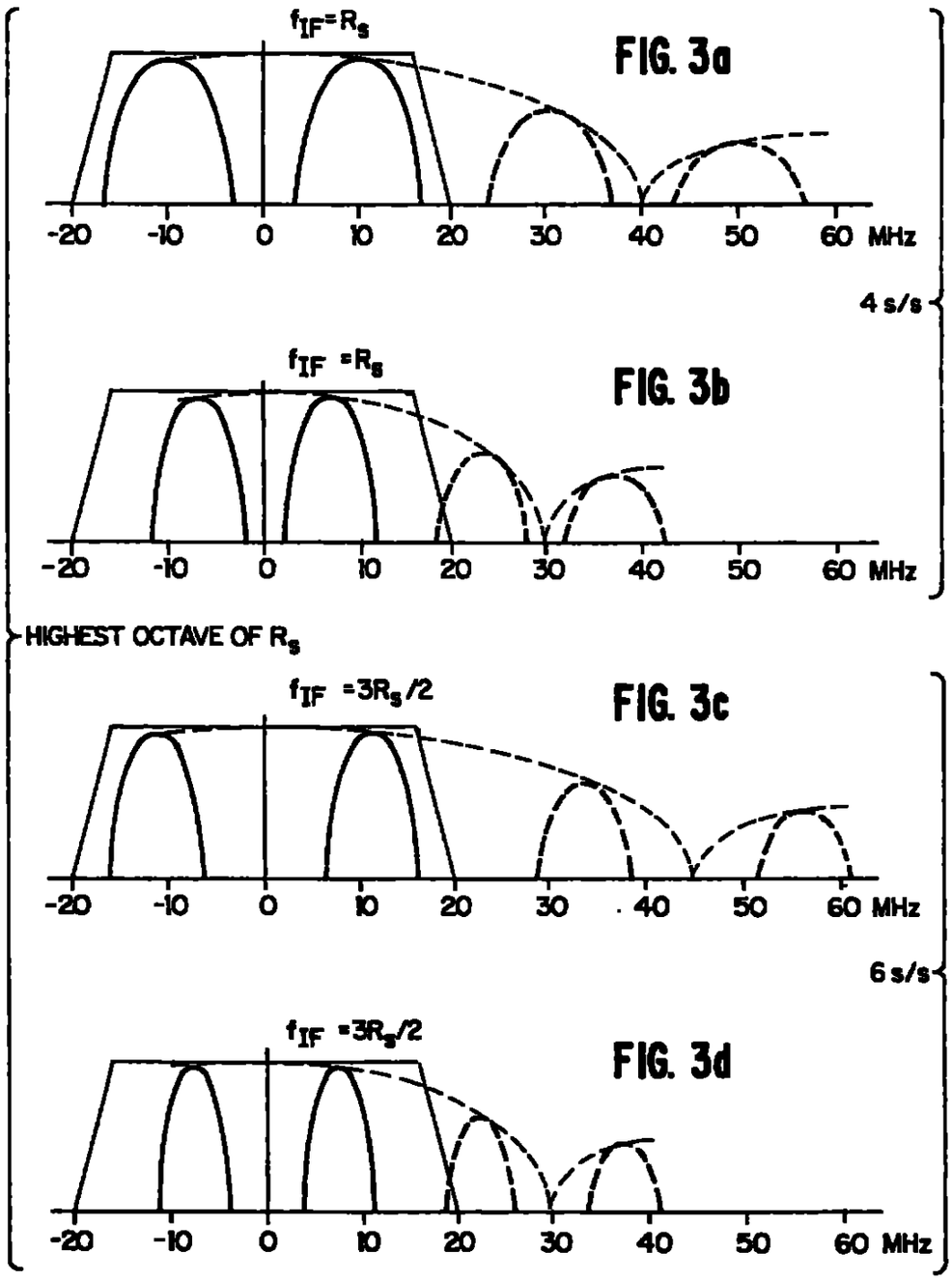


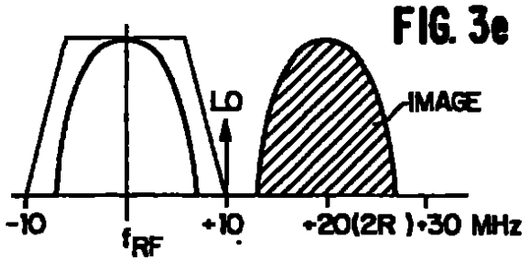
210

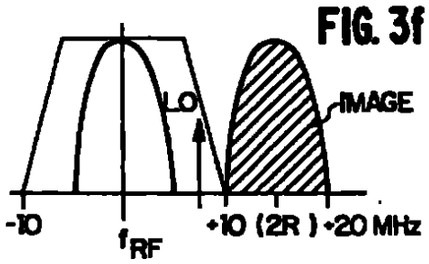
FIG. 2b
PRIOR ART



203







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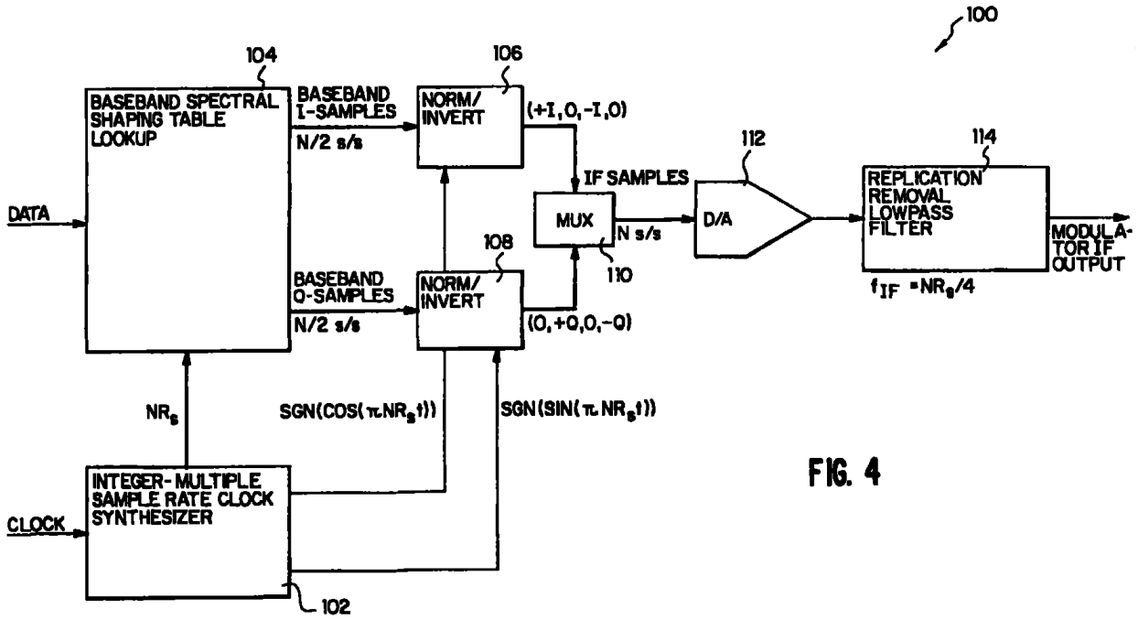

100
104
BASEBAND SPECTRAL SHAPING TABLE LOOKUP
DATA
BASEBAND I-SAMPLES
N/2 s/s
BASEBAND Q-SAMPLES
N/2 s/s
106
NORM/ INVERT
(+I,0,-I,0)
108
NORM/ INVERT
(0,+Q,0,-Q)
IF SAMPLES
MUX
N s/s
110
112
D/A
114
REPLICATION REMOVAL LOWPASS FILTER
MODULATOR IF OUTPUT
$f_{IF} = NR_s/4$
NR_s
SGN(COS(π.NR_st))
SGN(SIN(π.NR_st))
INTEGER-MULTIPLE SAMPLE RATE CLOCK SYNTHESIZER
CLOCK
102

FIG. 4

FIG. 5a

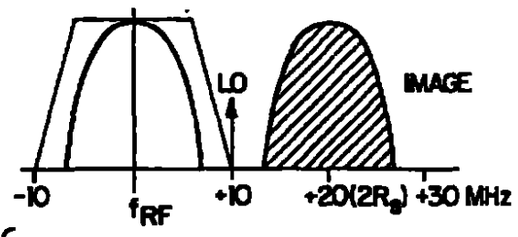


FIG. 5b

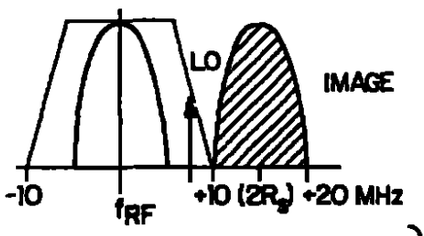


FIG. 5c

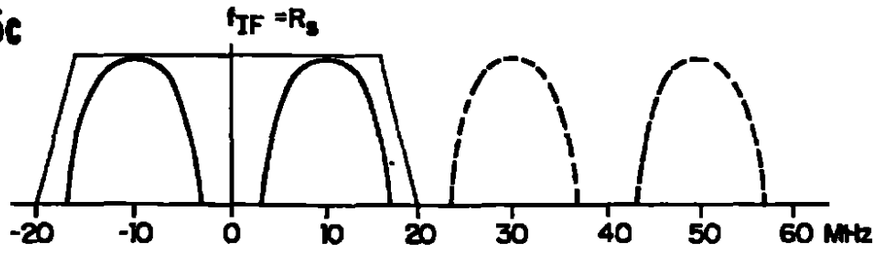
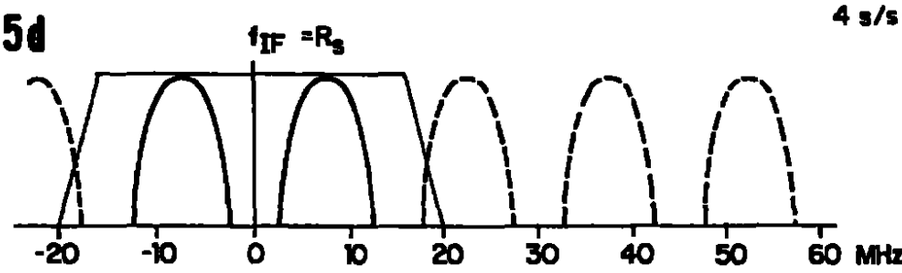


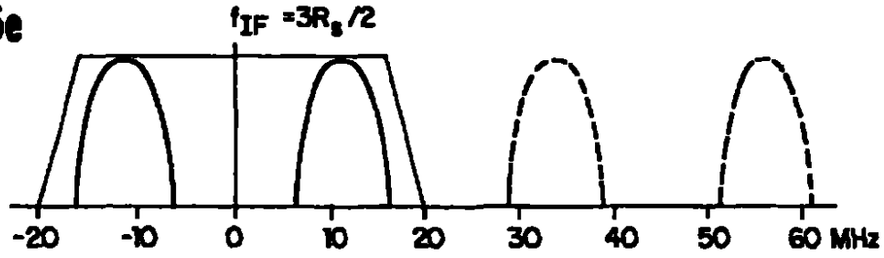
FIG. 5d



4 s/s

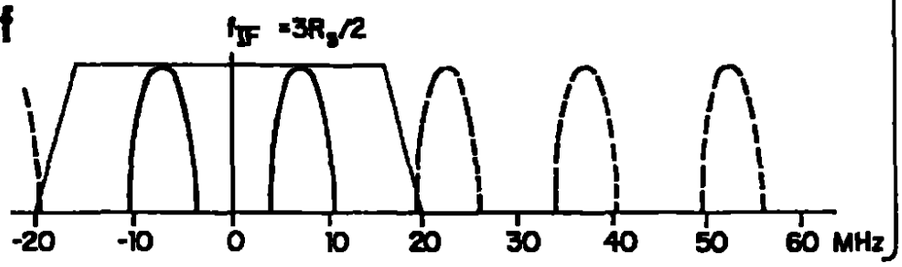
HIGHEST OCTAVE OF R_s

FIG. 5e



6 s/s

FIG. 5f



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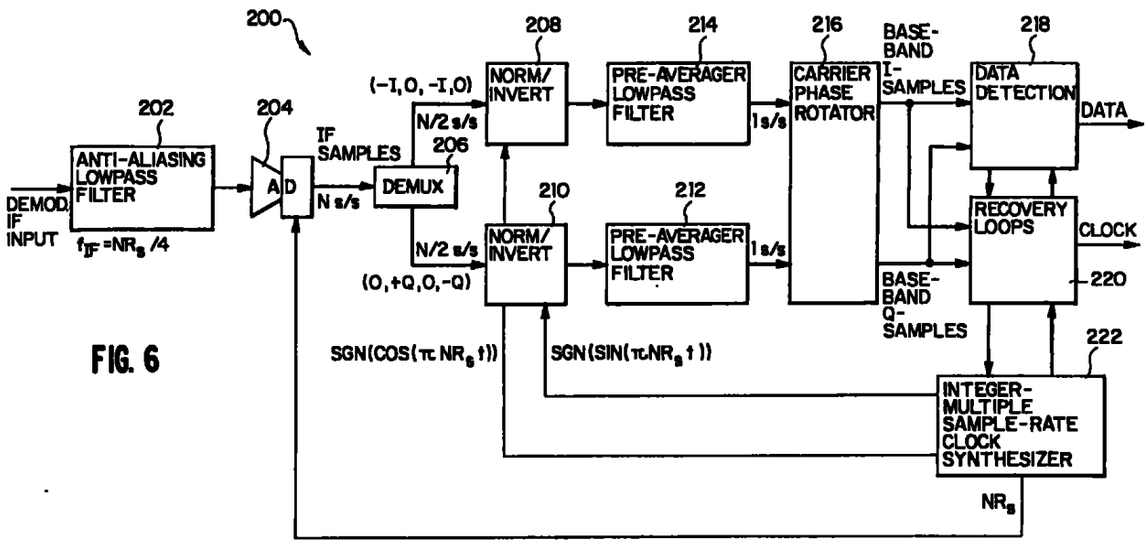

200
202
ANTI-ALIASING LOWPASS FILTER
DEMOD. IF INPUT
$f_{IF} = NR_s/4$
204
A/D
N s/s
IF SAMPLES
206
DEMUX
N/2 s/s
(-I, 0, -I, 0)
N/2 s/s
(0, +Q, 0, -Q)
208
NORM/ INVERT
210
NORM/ INVERT
212
PRE-AVERAGER LOWPASS FILTER
214
PRE-AVERAGER LOWPASS FILTER
1 s/s
1 s/s
216
CARRIER PHASE ROTATOR
BASE-BAND I-SAMPLES
BASE-BAND Q-SAMPLES
218
DATA DETECTION
DATA
RECOVERY LOOPS
CLOCK
220
222
INTEGER-MULTIPLE SAMPLE-RATE CLOCK SYNTHESIZER
NR_s
SGN(COS(τc NR_s t))
SGN(SIN(τc NR_s t))

FIG. 6

CONTINUOUSLY VARIABLE IF SAMPLING METHOD FOR DIGITAL DATA TRANSMISSION

BACKGROUND OF THE INVENTION

1. Field of the Invention

The present invention relates generally to modulation and demodulation of signals in variable rate information transmission systems. More specifically, the present invention relates to a sampling method for continuously sampling the intermediate frequency associated with digital data transmission.

2. Brief Discussion of Related Art

Modulated signals carry information such as video, data, music and speech. Efficient demodulation requires distinguishing the information from the attendant carrier wave. A general demodulation process includes several steps. The receiver may receive, at its antenna, an information signal modulated on a radio frequency (RF) carrier wave. The signal may then undergo frequency conversion to the intermediate frequency (IF) band. The information signal, at baseband, is recovered from the IF signal by a suitable detector. Considering, for example, a conventional receiver in a variable rate digital data transmission system, the IF signal, produced from a received RF signal by subjecting the RF signal to a mixing or filtering process, is subsequently applied to a data detector for recovering, at baseband, the information content of the input signal.

There are fundamentally two types of practical sampling techniques or methods: baseband sampling and IF sampling. To generate or recover an arbitrary bandpass signal with baseband sampling, the information signal must be frequency translated between an IF carrier frequency and baseband in the analog domain as shown in FIGS. 1A and 1B. In-phase and quadrature (I and Q) D/A or A/D converters are used to transfer the samples between the digital and analog domains.

It will be noted that any phase imbalance in the analog quadrature frequency translation causes crosstalk between the channels. Carrier feedthrough in the analog mixing process results in DC offsets in both the I and Q channels. It will also be noted that the amplitude needs to be balanced between the channels. In terms of the data symbol rate (R_s), the minimum integer-multiple baseband sample rate that can be used in each parallel I and Q channel is $2R_s$, or two samples-per-symbol (s/s).

With IF sampling, the bandpass signal is sampled directly, so only a single D/A or A/D converter and attendant bandpass filter are needed, as illustrated in FIGS. 2A and 2B. The digital IF samples are typically translated to baseband for processing. It is a trivial matter to set up this translation in the digital domain so there is no carrier frequency phase imbalance, hence, no I/Q crosstalk. For example, with the IF centered at R_s and sampled at 4 s/s, the quadrature translation can be implemented through multiplication by cosines and sines whose sample values are (+1, 0, −1, 0) and (0, +1, 0, −1), respectively. Furthermore, because of the digital translation there are no DC offsets or amplitude balance impairments.

It will be noted that the system must be responsive to a variable rate signal. Thus, the IF bandwidth must be broad enough to process the highest expected data rate, although at any point in time the receiver may be detecting a lower rate and thus a narrower band signal.

It will be also be recognized that the primary disadvantage associated with IF sampling is that the sample rate at the D/A's and A/D's must typically be higher than the $2R_s$ sampling rate required at baseband to avoid aliasing of the IF spectrum. It must be appreciated that in baseband sampling, the aggregate parallel sample rate of the I and Q channels is actually $4R_s$. Thus, the only major disadvantage with IF sampling is that the D/A and A/D converters must run faster.

U.S. Pat. No. 5,130,802 relates to a television receiving section, and discloses a manner of holding the IF to an integral multiple of the sample clock frequency of the A/D converter to thereby avoid unwanted signal folding or aliasing, wherein the intermediate frequency and the sample clock frequency are fixed depending on boundary conditions.

U.S. Pat. No. 4,859,960 discloses a method of demodulating a signal including a high frequency carrier signal having at least one parameter modulated by a low frequency modulating signal. The method includes translating the signal to a low intermediate frequency and digitizing the translated intermediate frequency signal at a correspondingly low sampling rate. On the other hand, U.S. Pat. No. 4,896,336 achieves symbol synchronization independent of the carrier phase or frequency.

SUMMARY OF THE INVENTION

The principal purpose of the present invention is to provide a simple and efficient means of performing intermediate frequency (IF) sampling in modulators and demodulators for digital data transmission over a broad range of data rate operations. According to one aspect of the invention, the effective range is limited only by the resolution and speed of the D/A and A/D converters.

An object of the of the present invention is to provide a digital transmission and reception system accommodating variable symbol rates wherein only a single low pass filter having a fixed bandwidth can be used at each end of the transmission path. According to one aspect of the invention, only a single analog low pass fixed bandwidth replication-removal filter is provided in the transmission signal path of the transmitter. According to another aspect of the invention, only a single analog low pass fixed bandwidth anti-aliasing filter is provided in the receiver.

These and other objects, features and advantages according to the present invention are provided by a sampling method, denoted continuously variable IF sampling (CVIFS), implemented in digital transmission and reception components. Advantageously, the method can be implemented to achieve these objectives without any hardware changes, i.e., the sampling rate of operation is established simply by programming the associated digital hardware. The use of the CVIFS method allows sampling of a bandpass signal which is centered at an intermediate frequency f_{IF}, that is related to both the data transmission symbol rate (R_s) and sample rate (NR_s) by integer ratios. In this manner, the implementationally cumbersome problem of sample interpolation is avoided.

These and other objects, features and advantages according to the present invention are provided by a CVIFS method characterized in that the sample rate NR_s is constrained to be four times the IF center frequency, and both the sample rate NR_s and the IF center frequency are constrained to be within a range of a factor of 1.5 of their maximum values. To maintain this high sample rate as the data symbol rate is lowered, a "gear shifting" technique is utilized whereby the integer number of samples-per-symbol, N, is roughly proportionally increased. Moreover, the sample rate NR_s is

always between 3.0 to 4.5 times the maximum data symbol rate (R_{max}), and the IF center frequency is always between 3/4 to 9/8 times (R_{max}), irrespective of the actual data rate R_s.

These and other objects, features and advantages of the invention are disclosed in or apparent from the following description of preferred embodiments.

BRIEF DESCRIPTION OF THE DRAWINGS

The preferred embodiments are described with reference to the drawings in which like elements are denoted by like or similar numbers and in which:

FIGS. 1A and 1B are high-level block diagrams of a baseband sampled modulator and baseband sampled demodulator, respectively, using conventional transmitters and receivers;

FIGS. 2A and 2B are high-level block diagrams of intermediate frequency (IF) sampled modulator and demodulator systems for respective transmitter and receiver components;

FIGS. 3A through 3F are illustrative waveforms which are useful in explaining the operation of continuously variable IF sampling in a transmitter according to the present invention;

FIG. 4 is a high-level block diagram of a continuous variable IF sampling modulator for a transmitter according to the present invention;

FIGS. 5A through 5F are illustrative waveforms which are useful in understanding the operation of a receiver implementing continuously variable information IF sampling according to the present invention; and

FIG. 6 is a high-level block diagram of a continuous variable IF sampling demodulator used in a receiver according to the present invention.

DESCRIPTION OF THE PREFERRED EMBODIMENTS

When a single-rate, fixed bandwidth analog signal is to be IF sampled and processed digitally, numerous well-known techniques could be applied, which techniques do not address the more general problem of digitally processing a continuously variable rate signal in an efficient manner. For example, as described above, a hardware efficient IF sampling technique for processing a fixed rate signal is to set the IF frequency equal to the data symbol rate and sample at four samples per symbol, or equivalently four times the IF carrier frequency. This technique greatly simplifies the subsequent processing, as conversion of the samples from IF to baseband merely entails multiplication by ±1 and 0. In addition, because of the multiplications by zero, the sampling rate can immediately be decimated by a factor-of-two in both the cosine and sine quadrature channels. However, when the data rate is variable over a relatively broad range, the IF frequency will change, and the analog filters required on the transmit side for sample clock replication removal (RR) as well as the receive side for anti-aliasing (AA) must be modified, e.g., scaled. It will be appreciated that single fixed bandwidth RR and AA filters are desirable.

The digital processing required for data transmission on both the transmit and receives sides advantageously can be greatly simplified when an integer number of s/s is used, where some of the receive samples are phase coherent with the data detection and transition points. With non-integer numbers of s/s, interpolating filters are needed to convert to integer s/s. Moreover, not only are these filters very hardware intensive but they also substantially increase the complexity of the sample processing. As discussed above, processing can be further simplified if the data sample rate is four times the IF center frequency, which is advantageously an integer ratio multiple of the data symbol rate. It will be appreciated that an integer relationship between the data symbol rate, sample rate, and the IF center frequency is desirable.

The CVIFS technique avoids all of the disadvantages associated with previous baseband and IF sampling methods with the single exception of increasing the maximum rate at which the D/A and A/D converters must operate. A better understanding of the CVIFS technique can be had by examining the maximum and minimum symbol rate extremes at different numbers of samples-per-symbol, N, for both the modulator and demodulator, thereby demonstrating continuously variable rate operation. Preferably, there will always be two integer values of N within each octave of R_s, that is, N=(4,6); (8,12); (16,24); . . .; for octaves 1,2,3, . . . This technique of bracketing N in each octave of R_s will hereafter be referred to as "gear shifting". A baseband sampling version of this gear shifting scheme was discussed in a paper by the instant inventor entitled "Programmable Digital Modem", presented at the NASA Space Communications Technology in November, 1991, which is incorporated herein for all purposes.

A conceptual diagram of the desired modulated signal spectral range is given in FIGS. 3A and 3B. As shown in these figures, R_{max} has been set to 10 Msymbol/s for the purpose of illustration. It will be appreciated that a fixed RR low pass filter outline is situated between −20 and +20 MHz. Preferably, this filter has a single-sided bandwidth of 1.6 R_{max} and a 30 dB shape factor of 1.25: The sin(x)/(x) rolloff envelope of the D/A converter is also indicated by the dashed lines in FIGS. 3A and 3B. Advantageously, all of the frames reflect modulated bandwidth spectral occupancies of 1.4 R_s, although it should be noted that this is not a restriction. The subject IF sampling scheme should perform well over the entire range of Nyquist filtered spectral occupancies, R_s to $2R_s$.

FIG. 3A corresponds to the desired IF output when the sample rate, NR_s, is $4R_{max}$ and the IF center frequency, f_{IF}, is R_{max}. As will soon become apparent, the ratio of these two quantities will always be 4. It will be appreciated that the main spectral lobe at the minimum symbol rate is easily contained within the RR filter's bandwidth, and the replicated spectra centered at integer multiples of the sample clock are readily attenuated. The minimum symbol rate to be used with 4 s/s is 7.5Msymbols/s as shown in FIG. 3B. Again, it will be appreciated that the IF center frequency has slid down and the RR filter is just able to reject the first replicated lobe. These exemplary cases represent the extremes of operation at N=4 s/s.

For the range of $5.0<R_s<7.5$Msymbols/s, N is set equal to 6 s/s, as shown with respect to FIGS. 3C and 3D, wherein operation at the extremes of this range are depicted. It should be appreciated from inspection of these figures that RR filtering becomes progressively less critical in terms of passing the main lobe and rejecting the replications as R_s is continually lowered. Moreover, the IF center frequency and sample rate advantageously are equal to $NR_s/4$ and NR_s for all rates of operation. Preferably, these values are always bounded by ($1.125\ R_{max} \geq f_{IF} > 0.75\ R_{max}$) and ($4.5\ R_{max} \geq NR_s > 3.0\ R_{max}$), respectively.

The relationships between N, R_s, f_{IF}, and NR_s are illustrated in Table 1, below, for R_{max}=10Msymbol/s. While only 3 octaves of R_s are listed in Table 1, the pattern for

subsequent octaves is easily extrapolated from the information provided.

TABLE 1

Modulator Gear Shifting (R_{max} = 10 Megasymbols)

Samples/ Symbol (n/s)	Symbol Rate (Megasymbols/s)	IF Center Freq (MHz)	Sample Rate (Megasamples/s)
N = 4	$10 \geq R_s > 7.5$	$10 \geq f_{IF} > 7.5$	$40 \geq NR_s > 30$
N = 6	$7.5 \geq R_s > 5.0$	$11.25 \geq f_{IF} > 7.5$	$45 \geq NR_s > 30$
N = 8	$5.0 \geq R_s > 3.75$	$10 \geq f_{IF} > 7.5$	$40 \geq NR_s > 30$
N = 12	$3.75 \geq R_s > 2.5$	$11.5 \geq f_{IF} > 7.5$	$45 \geq NR_s > 30$
N = 16	$2.5 \geq R_s > 1.875$	$10 \geq f_{IF} > 7.5$	$40 \geq NR_s > 30$
N = 24	$1.875 \geq R_s > 1.25$	$11.25 \geq f_{IF} > 7.5$	$45 \geq NR_s > 30$

From Table 1 it can be seen that in order not to restrict the maximum sample processing rate, the IF frequency must remain relatively low. Consequently, a second frequency translation will typically be needed. However, the lower the IF center frequency, the more difficult it is to separate the image band resulting from the second mixing process.

The advantage of keeping the IF frequency in a bounded range in the upper region of the RR filter's pass band becomes readily apparent by inspecting FIGS. 3B and 3F. It is clear from these figure that both the image and the local oscillator (LO) feedthrough can easily be separated with a bandpass filter at the worst case extremes of operation shown, provided that the filter characteristic is sufficiently narrow. In an exemplary case, the ratio of stop band bandwidth to center frequency in a practical crystal filter can be as low as 0.2% at center frequencies from 10 kHz to 100 MHz. Advantageously, the maximum symbol rate can be as small as the RF center frequency divided by 5×10^3 (f_{RF}=10 MHz, R_{max}=20 KHz).

FIG. 4 is a high level block diagram of an exemplary digital data transmission modulator **100** implementation with IF sampling according to the CVIFS technique according to the present invention. Modulator **100** advantageously includes an integer-multiple sample rate clock synthesizer **102**, connected to both a baseband spectra shaping lookup table **104** and normal/inverting circuits **106**, **108**, the outputs of which are provided to digital-to-analog (D/A) converter **112** via multiplexer **110**. Advantageously, the output of D/A converter **112** is provided to RR lowpass filter **114**.

It will be appreciated that an integer multiple sample rate clock must be synthesized from the incoming data clock. Moreover, in order to generate arbitrary bandpass signal spectra, two quadrature baseband modulating channels are necessary, designated I and Q. For simplicity, the incoming data is mapped into equivalent I and Q baseband samples. The baseband samples are translated up to the IF frequency by multiplication with cosine and sine samples at the IF frequency. Because the sample rate is always 4 f_{IF}, the cosine and sine samples are (+1, 0, −1, 0) and (0, +1, 0, −1), respectively. Hence, the I and Q channels need only be sampled at half the IF sample rate for all possible data rates of operation. The ±1 multiplication can be realized with simple normal/invert gating circuits **106**, **108** with one's complement or sign and magnitude arithmetic. Advantageously, because of the staggered multiplication by zero, the modulated samples can simply be multiplexed together into a single IF stream, using multiplexer **110**, rather than being summed. The D/A converter **112** converts the samples into the analog domain and imparts a sin(x)/(x) envelope on their spectra due to its inherent sample-and-hold effect. Finally, a single replication-removal low pass filter **113** advantageously can be used to eliminate spectral replications at multiples of the sample clock frequency, thus and the resulting IF output is centered at $NR_s/4$.

A conceptual diagram of the demodulated signal spectra is shown in FIGS. 5A–5F. From inspection of these figures, as well as FIG. 6, it will be appreciated that the sample-and-hold (S&H) aperture in the receive A/D **204**, which is described in greater detail below, advantageously can be typically narrow enough that its sin(x)/(x) rolloff effects can be ignored. It will also be appreciated that the sample clock replicated spectra are not significantly attenuated. The AA filter **202** has the same bandwidth and shape factor as the RR filter **114** used on the modulator **100**. However, its stop band attenuation preferably is increased to about 40 dB, as the sample and hold effect doesn't provide much residual attenuation.

It will be appreciated that the image problem depicted in FIG. 5A is substantially identical to FIG. 3E regarding modulator **100** and that the same image reject filter could be used. Moreover, it should be noted that all of the pertinent sampling parameters (N, R_s, NR_s, and f_{IF}) scale in exactly the same way for demodulator **200** as they did for modulator **100**. For example, the sample rate is always four times the IF center frequency, the sample rate and IF frequency are constrained to be within a range of 1.5, and each octave of operation is broken up into two regions with different integer numbers of s/s. An additional advantage which can be attributed to the CVIFS technique is the commonality of the transmit and receiver IF frequencies, their image, RR, and AA filters, their maximum sample rates, as well as their sample rate gear shifting.

The high level block diagram for demodulator **200** in FIG. 6 has much of the same functionality as that of modulator **100**, albeit in reverse order. In addition, however, the demodulator data must be filtered and detected; the incoming signal's amplitude, carrier and symbol timing phase must be recovered and tracked.

Advantageously, the incoming analog signal will be centered at an IF frequency of $NR_s/4$ and, in an exemplary case, will be is passed through a low pass AA filter **202** is followed by an A/D converter **204**, which is sampled at NR_s. It will be appreciated that the sample rate is demultiplexed into I and Q channels by demultiplexer **206** and decimated by a factor-of-two prior to baseband translation. Because of the ±1 multiplication, only simple normal/invert gating circuitry **208**, **210** is required. With the samples now at baseband, the CVIFS technique can be used in conjunction with U.S. Pat. Nos. 5,052,027 and 5,216,696, which patents are incorporated herein by reference for all purposes, which disclose a pre-averaging technique used to reduce the sample rate for subsequent processing to 1 s/s.

In general, carrier frequency translation in the transmission link will result in an IF frequency offset, thereby causing a beat note in the I and Q channels. The beat note is easily removed with a carrier phase rotator **216**, which performs a 2×2 matrix multiplication of the baseband I and Q channels with cosines and sines of the carrier phase tracking loop output. Typically, beat notes as large as 10% of the data symbol rate can be removed. The CVIFS technique advantageously cannot be said to restrict this range. Since the sample rate is always an integer multiple of the symbol rate, the sample rate timing clock, NR_s, is actually recovered in the clock tracking loop **220**, and the symbol timing clock, r_s, is obtained through division by N for output from clock synthesizer **222**. The normal/invert IF frequency translation signals are procured with respect to the sample clock divided by two.

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It will be appreciated that the above described modulator and demodulator can be provided by an IF sampling mechanism accommodating a continuously variable rate operation by software changes; thus the preferred embodiments of the modulator and demodulator can be obtained without any hardware changes. Moreover, it will be appreciated that the gear shifting permits maximum image separation of the modulator and demodulator carrier IF frequencies, which results in a reduction in the complexity of the image rejection filter. Additionally, the use of gear shifting permits maintenance of the sample rate at a factor of four times the IF frequency over the entire range of symbol rates to be processed.

Other modifications and variations to the invention will be apparent to those skilled in the art from the foregoing disclosure and teachings.

In an exemplary case representing yet another embodiment of the CVIFS technique, the overall sample rate would be doubled, i.e., sampling at 8 s/s for the highest symbol rate of operation while centering the IF frequency at $2R_s$. Advantageously, this has the effect of permitting easier image rejection, albeit at the expense of reducing the maximum rate of operation by half.

It will be appreciated that RR and AA filters 114 and 202, respectively, with slightly different characteristics can advantageously be used, e.g., the filters could be alternating current (AC) coupled which, it will be readily noted, would render the RR and AA filters bandpass filters.

Advantageously, the CVIFS approach permits an extremely broad data rate range of operation using sample analog RR and AA filtering without additional costly or complex hardware changes, permits elimination of the interpolating filters often associated with other systems, while eliminating quadrature amplitude and phase balance as well as the DC offset adjustments commonly associated with IF-to-baseband conversion in analog front-ends with baseband sampling. Additionally, only a single D/A and A/D converter and analog filter are required on the transmit and receive sides, respectively.

It will also be appreciated that regardless of the data rate of operation, the conversion between the baseband and IF samples is always carried out at the same 4 samples-per-IF carrier cycle, which advantageously can be implemented using simple and robust normal/invert and multiplexing circuitry or software equivalents. Preferably, the sample rate is also decimated by a factor-of-two immediately prior to the D/A and immediately after the A/D conversion for all of the subsequent processing. Using the CVIFS technique, the requirements regarding image the rejection filter are simplified because the desired band and image band are maintained as far apart as possible. The CVIFS technique also permits the spectral occupancy of the IF sampled signal to be spread over the entire Nyquist band, R_s to $2R_s$.

Other modifications and variations to the invention will be apparent to those skilled in the art from the foregoing disclosure and teachings. Thus, while only certain embodiments of the invention have been specifically described herein, it will be apparent that numerous modifications may be made thereto without departing from the spirit and scope of the invention.

What is claimed is:

1. An intermediate frequency (IF) sampling method for continuously sampling the intermediate frequency associated with digital data transmission in a communications channel, said method comprising the steps of:

sampling a signal centered on an intermediate frequency f_{IF} which is related to both data transmission rate (R_s) and sample rate (NR_s) by an integer ratio; and

varying an integer number N of samples per symbol inversely with respect to a change is data symbol rate while maintaining said sampling rate at least equal to a Nyquist rate.

2. The intermediate frequency (IF) sampling method as recited in claim 1, further comprising the steps of constraining said sample rate NR_s to be four times an IF center frequency and constraining both said sample rate NR_s and said IF center frequency to be within a range between 1.5 times and four times respective maximum values of said sample rate NR_s and said IF center frequency.

3. The intermediate frequency (IF) sampling method as recited in claim 1, further comprising the steps of constraining said sample rate NR_s to be within a range of 3.0 to 4.5 times a maximum data symbol rate R_{SMAX} and constraining both said IF center frequency to be always be between 3/4 and 9/8 times said maximum data symbol rate R_{SMAX}.

4. An intermediate frequency (IF) sampling apparatus for continuously sampling the intermediate frequency associated with digital data transmission in a communications channel, said apparatus comprising:

means for sampling a signal centered on an intermediate frequency f_{IF} which is related to both data transmission rate (R_s) and sample rate (NR_s) by an integer ratio; and

means for varying an integer number N of samples per symbol inversely with respect to a change is data symbol rate while maintaining said sampling rate at least equal to a Nyquist rate.

5. The intermediate frequency (IF) sampling apparatus as recited in claim 4, further comprising:

means for constraining said sample rate NR_s to be four times an IF center frequency; and

means for constraining both said sample rate NR_s and said IF center frequency to be within a range between 1.5 times and four times respective maximum values of said sample rate NR_s and said IF center frequency.

6. The intermediate frequency (IF) sampling apparatus as recited in claim 4, further comprising:

means for constraining said sample rate NR_s to be within a range of 3.0 to 4.5 times a maximum data symbol rate R_{SMAX}; and

means for constraining both said IF center frequency to be always be between 3/4 and 9/8 times said maximum data symbol rate R_{SMAX}.

* * * * *

BNSDOCID <US 5696796A1 I >

215

UNITED STATES PATENT AND TRADEMARK OFFICE

CERTIFICATE OF CORRECTION

214

PATENT NO. : 5,696,798
DATED : December 9, 1997
INVENTOR(S) : John James Poklemba

It is certified that error appears in the above-identified patent and that said Letters Patent is hereby corrected as shown below:

The title page, showing an illustrative figure, should be deleted and substitute therefor the attached title page.

Delete Drawing Sheet 8 and substitute therefor the Drawing Sheet, consisting of FIG. 6, as shown on the attached page.

Signed and Sealed this

Ninth Day of February, 1999

Attest:

Attesting Officer

Acting Commissioner of Patents and Trademarks

BNSDOCID <US 5696798A1 >

United States Patent [19]

Poklemba

[11] Patent Number: 5,696,796

[45] Date of Patent: Dec. 9, 1997

[54] **CONTINUOUSLY VARIABLE IF SAMPLING METHOD FOR DIGITAL DATA TRANSMISSION**

[75] Inventor: John James Poklemba, Ijamsville, Md.

[73] Assignee: Comsat Corporation, Bethesda, Md.

[21] Appl. No.: 483,764

[22] Filed: Jun. 7, 1995

[51] Int. Cl.[6] H03K 9/00; H04L 27/06; H04L 27/14; H04L 27/22

[52] U.S. Cl. 375/316; 332/103

[58] Field of Search 375/316, 326; 332/103; 329/129

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Primary Examiner—Stephen Chin

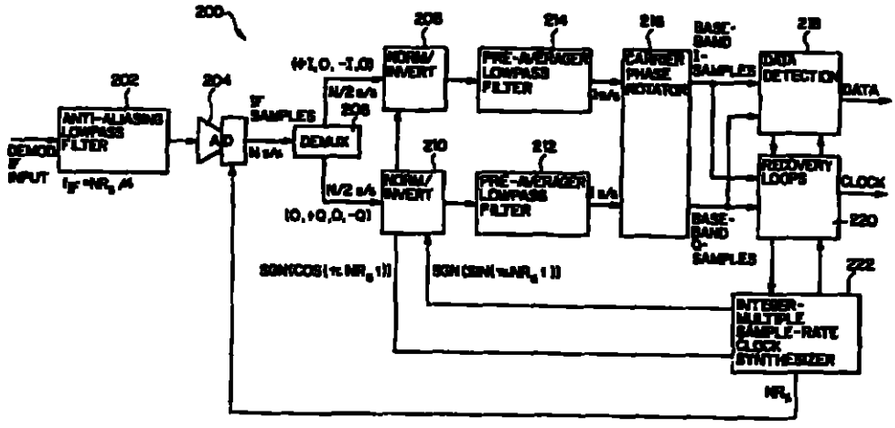
Assistant Examiner—Mohammad Ghayour

Attorney, Agent, or Firm—Sughrue, Mion, Zinn, Macpeak & Seas, PLLC

[57] ABSTRACT

A sampling method, denoted continuously variable IF sampling (CVIFS), implemented in digital transmission and reception components is characterized by sampling a bandpass signal which is centered at an intermediate frequency f_{IF} that is related to both the data transmission symbol rate (R_s) and sample rate (NR_s) by integer ratios. Moreover, the CVIFS method is characterized in that the sample rate NR_s is constrained to be four times the IF center frequency, and both the sample rate NR_s and the IF center frequency are constrained to be within a range of a factor of 1.5 of their maximum values. To maintain this high sample rate as the data symbol rate is lowered, a "gear shifting" technique is utilized whereby the integer number of samples-per-symbol, N, is increased as the symbol rate decreases. Moreover, the sample rate NR_s is always between 3.0 to 4.5 times the maximum data symbol rate ($R_{s_{max}}$), and the IF center frequency is always between 3/4 to 9/8 times ($R_{s_{max}}$), irrespective of the actual data rate R_s.

6 Claims, 8 Drawing Sheets



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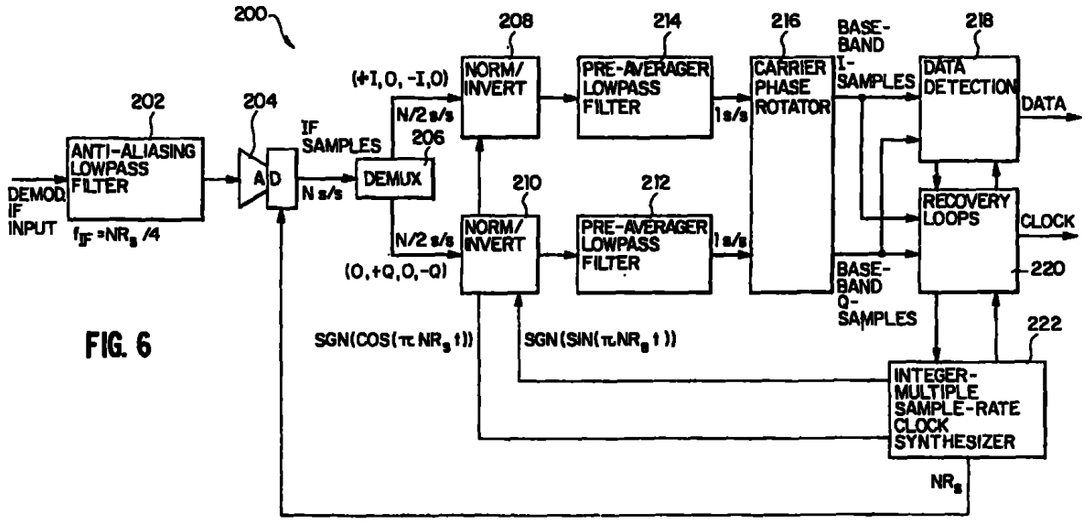

200
202
ANTI-ALIASING LOWPASS FILTER
DEMOD. IF INPUT
$f_{IF} = NR_s/4$
204
A/D
N s/s
IF SAMPLES
DEMUX
206
N/2 s/s
(+I, 0, -I, 0)
N/2 s/s
(0, +Q, 0, -Q)
208
NORM/ INVERT
210
NORM/ INVERT
212
PRE-AVERAGER LOWPASS FILTER
214
PRE-AVERAGER LOWPASS FILTER
1 s/s
1 s/s
216
CARRIER PHASE ROTATOR
BASE-BAND I-SAMPLES
BASE-BAND Q-SAMPLES
218
DATA DETECTION
DATA
RECOVERY LOOPS
CLOCK
220
222
SGN(COS(π NR_s t))
SGN(SIN(π NR_s t))
INTEGER-MULTIPLE SAMPLE-RATE CLOCK SYNTHESIZER
NR_s

FIG. 6

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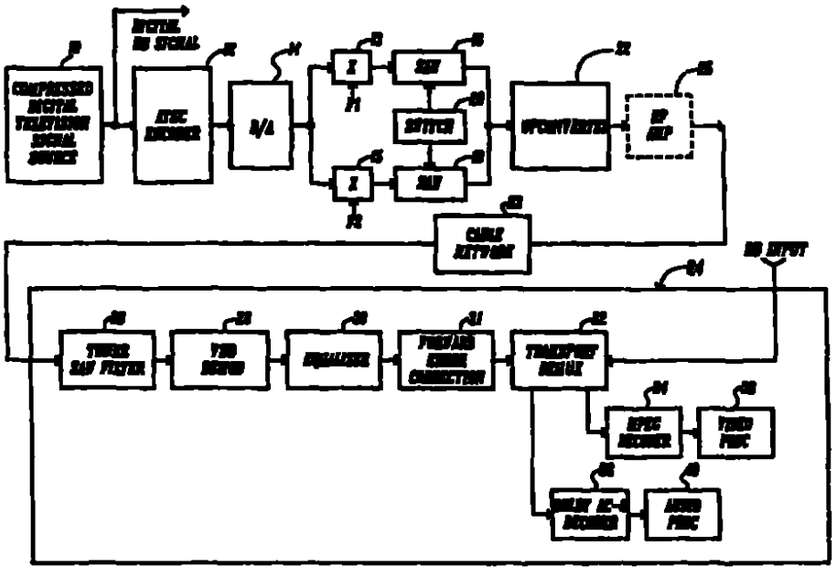
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(54) Title: VSB ENCODER AND RF MODULATOR FOR DIGITAL TELEVISION RECEIVER



(57) Abstract

An encoder–modulator (12, 13, 15) for coupling a digital baseband television signal to a VSB digital television receiver (24) that includes filters (26), equalization circuitry (30) and forward error correction circuitry (31) for correcting signal impairments that are below a given threshold. The encoder modulator processes the baseband signal for low power transmission on an RF channel with less–than–nominal bandpass characteristics. A coaxial cable (23) or other low noise communication link directly connects the RF signal to the RF input of the television receiver (24). Any errors or signal impairments in the transmitted signal that are below the predetermined threshold are corrected by the filters and equalization circuitry built into the VSB digital television receiver (24).

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EMSDOCID <WO 9940721A1 I >

VSB ENCODER AND RF MODULATOR FOR DIGITAL TELEVISION RECEIVER

Background of the Invention and Prior Art

This invention relates generally to digital VSB (vestigial side band) television receivers and specifically to a simple low cost system for coupling a digital VSB signal from any of a plurality of sources of MPEG (Motion Picture Experts Group) digital signal to a digital VSB television receiver. While the description is directed to a VSB digital format, it should be understood that the invention in its broader aspects is not limited to a particular digital format.

The recently adopted Digital Television Standard specifies a VSB subsystem having a terrestrial broadcast mode (8VSB) and a high data rate mode (16 VSB). There are other VSB modes available, namely 2 VSB, 4 VSB and 8 VSB (non terrestrial). The 8 VSB terrestrial mode has the data carrying capability of 4 VSB. (The Trellis coding that is added for the terrestrial environment creates the additional modulation levels.) The digital data signal is MPEG and Dolby AC-3 processed and must be subjected to decompression before application to conventional video and audio circuitry. For terrestrial broadcasting, the data signal is: randomized; subjected to Reed-Solomon (R/S) type encoding for error correction; interleaved; Trellis encoded; multiplexed with segment sync and field sync; supplied with a DC pilot; subject to pre-equalization filtering; modulated; and RF upconverted for transmission. The digital television receiver includes a tuner and a VSB demodulator for developing the baseband signal, which is in compressed form. The demodulated signal is applied to an equalizer for equalizing the received signal. The baseband signal is applied to a transport demultiplexer which directs the data to an appropriate MPEG decoder and a Dolby decoder for recovering the video and audio in proper form for application to video and audio processing circuitry. Digital signals from other sources, such as DVD (digital video disk) players, VCRs (video cassette recorders), PCs (personal computers), digital cable boxes, satellite receivers and the like will be supplied to

ENSDOCID <WO 9940721A1 I >

digital television receivers, as is the case for present analog television receivers.

In an analog system, it is well known that coupling the signal from a VCR or other video source to a baseband input of a television receiver (if one is available) yields superior results than those achieved by modulating the signal to RF channel 3 or channel 4 and supplying it to the tuner input of the television receiver. This is not, however, necessarily true in a digital environment in which case the distance between the baseband signal source and the television receiver must be kept relatively short.

In accordance with the invention, a compressed baseband digital signal is encoded in a given format for digital transmission and modulated on an RF carrier for direct application via a cable network or wireless link to the RF input of a digital television receiver. The RF signal, which is at a low power level and has a less-than-nominal bandpass, takes advantage of the front-end signal processing that is built into the digital television receiver. Signal impairments, due to the less-than-nominal bandpass characteristic and noise introduced during transmission, that are below a given threshold are corrected by the correction circuitry in the digital television receiver front end. The low signal power, the corrective circuitry for the digital signal and the benign environment of the communications link, enable components and circuits of much lower tolerance (and cost) to be used in the encoding and modulating process. The result is an encoder-modulator that is very low cost and very effective in coupling a digital baseband signal to a digital television receiver over an RF channel.

Objects of the Invention

A principal object of the invention is to provide a novel digital signal translation system.

Another object of the invention is to provide a digital encoder-modulator for coupling a baseband digital signal to a VSB digital television receiver.

A further object of the invention is to provide a low cost digital signal coupling system for a VSB digital television receiver.

Brief Description of the Drawings

These and other objects and advantages of the invention will be apparent upon reading the following description in conjunction with the drawing, in which:

ENSDOCID <WO 9940721A1 I >

FIG 1 is a simplified block diagram of a VSB digital signal encoder-modulator constructed in accordance with the invention:

FIG 2 is a simplified block diagram of another form of a digital signal encoder-modulator constructed in accordance with the invention: and

FIG 3 are curves illustrating the bandpass characteristics of the encoder modulator of the invention.

Description of the Preferred Embodiment

Referring to FIG 1, a source of digital MPEG encoded signal 10 derived from any of a number of different source types, is shown. For example, the source may comprise a satellite receiver, a VCR, a DVD, a digital cable box, a PC, etc. The output of source 10 is an MPEG (and Dolby AC-3) encoded digital signal at baseband frequency. This baseband signal may be applied directly to a baseband input of a digital television receiver 24. However, in accordance with the invention, the baseband signal is applied to an ATSC encoder 12 where the signal is subjected to interleaving, randomizing, R/S error encoding and Trellis encoding. Field and segment syncs are added, and while not indicated, a pilot is inserted and equalization filtering is employed. (See ATSC Standard A/53, ATSC Digital Television Standard for a detailed description of these operations.) The signal is then supplied to a digital to analog (D/A) converter 14. D/A 14 supplies the analog signal to a pair of multipliers 13 and 15 that in turn supply signals to a pair of SAW (surface acoustic wave) filters 16 and 18. Multipliers 13 and 15 are each supplied with a frequency F1 and F2, respectively for producing two different outputs. The SAW filters are selectively controlled by a switch 20, with one of the SAWs being used for one selected channel RF output and the other being used for another selected channel RF output. The particular RF channel selected is dependent upon the environment in which the encoder-modulator is used, in particular the availability and type of RF channels in the receiving area. The SAW filters are relatively low cost devices and are extensively used in television receivers.

Those skilled in the art will recognize that highly accurate modulators generate I and Q signals with the signals being combined to cancel one of the sidebands and

ENSDOCID <WO 9940721A1 I >

generate a vestigial sideband signal. Such modulators are quite complex and very expensive. In one form of the invention, that arrangement is replaced with a multiplier (13, 15)and a SAW filter (16, 18). The output of the multiplier is a double sideband suppressed carrier signal. The following SAW filter removes most of one sideband (preferably, although not necessarily, the lower sideband) to produce a VSB signal and also approximates the nominal Nyquist slope of the channel bandpass. As a result of the less than nominal bandpass characteristic, the transmitted signal will be characterized by a degree of impairment in the form of intersymbol interference (ISI).

The SAW filters shape the edges of the signal bandpass to approximate the Nyquist slopes, nominally called for at the band edges in the ATSC standard. (The Nyquist slopes may be seen by reference to the solid line curve of FIG 3.) The SAW filter characteristics also provides for adjacent channel rejection. The filtered signal from the selected one of the SAWs is applied to an RF upconverter 22, which upconverts the signal to any desired RF channel, channels 3 or 4 being the ones normally used.

The encoded and modulated VSB signal is supplied over a cable network 23, which may comprise a single coaxial cable or a relatively complex home cable network, to an RF channel input of a tuner 26 of a VSB digital television receiver 24. The signal may also be amplified by a low power RF amplifier 25 (shown in dotted line in FIG. 1) and transmitted wirelessly to the digital television receiver in an assigned RF broadcast television channel. The received signal is supplied through a companion SAW filter to a VSB demodulator 28 where the baseband digital signal is recovered. The demodulated signal is applied to an equalizer 30 that operates to adjust the response of the receiver to very closely match the solid line Nyquist response (FIG 3) thereby correcting any signal impairments, including any intersymbol interference introduced in the transmitter and any linear distortions introduced in the communications link (i.e. the cable network or wireless link) that collectively are below a given threshold level that is dependent upon the nature of the equalizer. The signal from equalizer 30 is processed by a forward error correction subsystem 31 to correct any residual errors due to intersymbol interference and errors due to noise,

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both white and impulse. introduced by the cable network or wireless link. The performance of equalizer 30 and error correction subsystem 31 sets the limit on the amount of intersymbol interference and linear distortion which can be introduced by the transmitter and communications channel before the system suffers serious degradation.

The corrected signal (in MPEG encoded form) is supplied to a transport demultiplexer 32 that has separate outputs for the video and audio portions of the signal. The video portion is applied to an MPEG decoder 34 for decompression and decoding and the audio portion is applied to a Dolby AC-3 decoder 36 for complementary processing. The resultant video and audio signals are applied to a video processor 38 and an audio processor 40, respectively.

It will be appreciated by those skilled in the art that the RF signal is not as distance sensitive as a baseband signal and the coaxial cable environment further minimizes errors or impairments being introduced in transmission. However, even if some signal impairments are introduced, they may be compensated for by the corrective circuitry built into the VSB digital television receiver, in particular the equalization circuitry. The result is a signal translation arrangement that results in excellent transfer of signal from source 10 to VSB digital television receivcr 24, since the signal is VSB encoded and any errors or signal impairments below a given threshold are correctable in the television receiver.

In contrast, any baseband digital signals that are supplied to television receiver 24 are introduced after equalizer 30 and error correction subsystem 31 and are consequently not subject to the corrective effects of the circuitry in the VSB digital television receiver. Such signals are also distance limited and subject to uncorrectable errors or impairments, which makes the method of the invention far superior.

In further accordance with the invention, where the digital television receiver has provision for receiving both a digital and an analog NTSC signal, in establishing priority of the RF channel output of upconverter 22, the first choice is to supply the signal on a vacant RF channel in the service area of the television receiver and the second choice is to supply it on the digital one of a pair of digital and analog NTSC

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RF channels in the service area of the television receiver. In this way, any NTSC interference into the digital channel can be minimized by the circuitry in the digital channel of the television receiver.

The invention resides in the concept of introducing a given level of impairments in the encoded digital baseband signal and taking advantage of the signal correction circuitry that is built into the front end of the digital television receiver to compensate for such impairments. The cost is kept minimal since in the application of the invention, transmission power is very low, distance is limited, the coaxial cable is a low noise environment and the channel shape (Nyquist slope) need not be as rigorously defined for adjacent channel rejection, signal radiation, etc. This translates into low cost filtering to create a less-than-nominal Nyquist slope. While this will clearly cause some intersymbol interference, a mentioned above, all such signal impairments that are below a predetermined threshold, may be compensated for in the digital receiver by the cooperative action of complementary filtering, the equalization circuitry, and the forward error correction circuitry.

The encoder-modulator of FIG 2 is a different version of the invention. Here the low cost SAW filters of FIG 1 have been replaced by digital filters 50, indicated as being FIR (finite impulse response) filters. A digital modulator 52 is supplied with the output of a first local oscillator 56 and converts the filtered input signal to a first IF signal having a frequency, for example, of about 12 MHz. This signal is applied to D/A 14 and thence to an RF upconverter 54 that is supplied with the output of a second local oscillator 58. Local oscillator 58 is controlled by a switch 60 to develop an RF output from upconverter 54 at either of a pair of RF channel frequencies, as in the tuner of FIG 1. The output signal is transmitted to the television receiver over coaxial cable 23. This version of the invention appears to be a more stable system. Since both systems will involve integrated circuitry, it remains to be seen whether the cost of the SAWs in the FIG 1 version will be less than the cost of the additional chip area required in the integrated circuit implementation of the FIG 2 version.

In FIG 3 the solid line curve indicates the ideal Nyquist slope channel bandpass response characteristic of the transmitted signal. The dashed line curve

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indicates the less-than-nominal Nyquist slope that results from using the low cost SAWs 16 and 18 in FIG 1 and the dotted line curve represents the response for the FIG 2 implementation using the FIR filters 50. The curves should be recognized as being representative only and their actual shapes are dependent upon the precision and the number of the filtering elements used. As mentioned, because of the less-than-nominal bandpass characteristic produced by either the SAW or the FIR filter embodiment of the invention, a certain level of intersymbol interference will be introduced into the transmitted signal. However, as discussed above, the amount of any such intersymbol interference, together with any noise introduced into the signal over the communications link, can be substantially corrected in the television receiver to provide near perfect signal translation and reception. In this manner the invention permits a low cost, high performance encoder-remodulator by deliberately selecting lesser tolerance components and applying the signal to a digital receiver that has the capability of compensating for signal impairments below a given threshold.

What has been described is a novel method and apparatus for translating a digital television signal from a digital source to a VSB digital television receiver. It is recognized that numerous changes to the described embodiment of the invention will be apparent to those skilled in the art without departing from its true spirit and scope. The invention is to be limited only as defined in the claims

BNSDOCID <WO 9940721A1 I >

What Is Claimed Is:

1. A method of operating a digital signal system for supplying a digital television receiver having correction circuitry capable of correcting impairments in a received signal that are below a

predetermined threshold comprising:

developing a compressed baseband digital signal;

encoding the compressed baseband signal in a given format for digital transmission; and

modulating the encoded compressed baseband digital signal on an RF carrier for direct application to the RF input of the digital television receiver;

the modulating step resulting in a less-than-nominal bandpass which produces impairments in the transmitted signal that are less than the predetermined threshold;

whereby the correction circuitry is capable of correcting the impairments introduced in the modulating step.

2. The method of claim 1, wherein the encoding step includes data interleaving, data randomizing, error protection and the addition of VSB field and segment syncs, and further comprising:

performing the modulating step at a low power level.

3. The method of claim 1, further comprising:

supplying the encoded compressed baseband digital signal through a first filter having a response designed to cooperate with a second filter in the digital television receiver for producing a less-than-nominal Nyquist slope.

4. The method of claim 1, further comprising:

supplying the encoded compressed baseband digital signal through a first filter having a response designed to cooperate with the equalization circuitry and a second filter in the digital television receiver for generating a nominal Nyquist slope and providing adjacent channel rejection at each end of the signal bandpass.

5. The method of claim 1, wherein the television receiver comprises a digital signal processing system and an analog NTSC signal processing system and further

BNSDOCID <WO 9940721A1_I >

comprising generating the modulated RF digital signal for application to the digital processing system of the television receiver.

6. The method of claim 1, wherein the RF carrier is applied to the RF input of the digital television receiver by a coaxial cable network.

7. The method of claim 1, wherein said modulating step comprises:

multiplying the baseband signal by a predetermined carrier frequency to generate a double sideband suppressed carrier signal; and

using a filter to remove one of the sidebands to develop a VSB signal and provide the less-than-nominal bandpass.

8. The method of claim 7, wherein the filter is a SAW and wherein the bandpass has a less-than-nominal Nyquist slope.

9. The method of claim 1, wherein the correction circuitry comprises equalization and forward error correction circuitry.

10. The method of claim 1, wherein the RF carrier is applied wirelessly to the RF input of the digital television receiver over a broadcast television channel.

11. A system for supplying a compressed digital baseband signal to a digital television receiver that includes correction circuitry capable of correcting impairments in a received signal that are below a predetermined threshold, comprising:

a source of compressed baseband signal;

an encoder for encoding said baseband signal in a given digital format;

a modulator for modulating said encoded baseband signal on an RF carrier;

said modulator having a less-than-nominal bandpass that produces impairments in the translated signal that are below said predetermined threshold; and

means for directly applying said modulated signal to an RF input of said digital television receiver, whereby said correction circuitry is capable of correcting said impairments introduced by said modulator.

12. The system of claim 11, wherein said encoder and said digital television receiver each include a filter, and wherein said filters cooperate to generate a less-than-nominal Nyquist slope.

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13. The system of claim 12, wherein said filters and said equalization circuitry cooperate to generate a nominal Nyquist slope and provide adjacent channel rejection at each end of the signal bandpass.

14. The system of claim 11, wherein said given digital format is VSB, said compressed digital signal is in MPEG form, said encoder includes a data randomizer, a data interleavor, error protection circuitry and means for adding VSB frame and field syncs and wherein said modulator operates at low power.

15. The system of claim 11, wherein said RF carrier is applied to the RF input of the digital television receiver by a coaxial cable network.

16. The system of claim 11, wherein said digital television receiver comprises a digital signal processing system and an analog NTSC signal processing system and wherein said modulated RF digital signal is generated for application to the digital processing system of the television receiver.

17. The system of claim 11, further comprising:

multiplying means in said modulator for developing a double sideband suppressed carrier signal from said baseband signal; and

filter means coupled to said multiplying means for substantially removing one of said sidebands to develop a VSB signal and for producing said less-than-nominal bandpass.

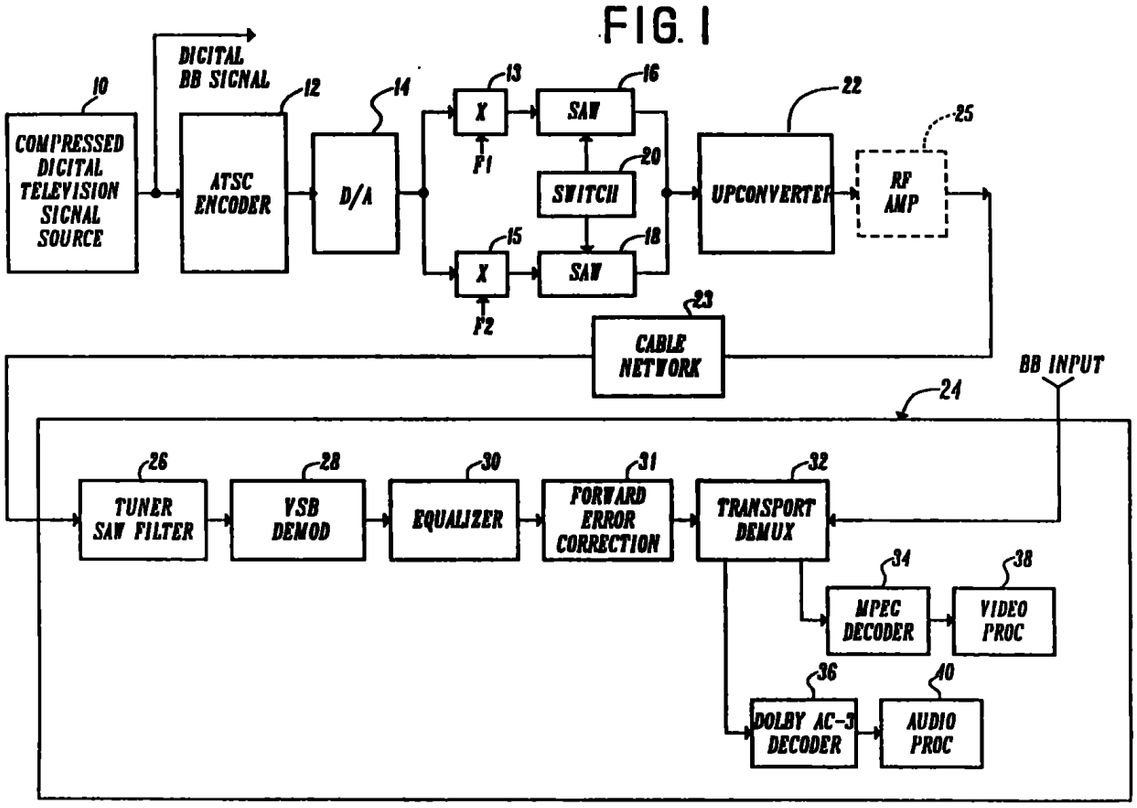
18. The system of claim 17, wherein said filter means comprises a SAW filter and wherein said bandpass includes a less-than-nominal Nyquist slope.

19. The system of claim 11, wherein the correction circuitry comprises equalization and forward error correction circuitry.

20. The system of claim 11, wherein the RF carrier is applied wirelessly to the RF input of the digital television receiver over a broadcast television channel.

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FIG. 1



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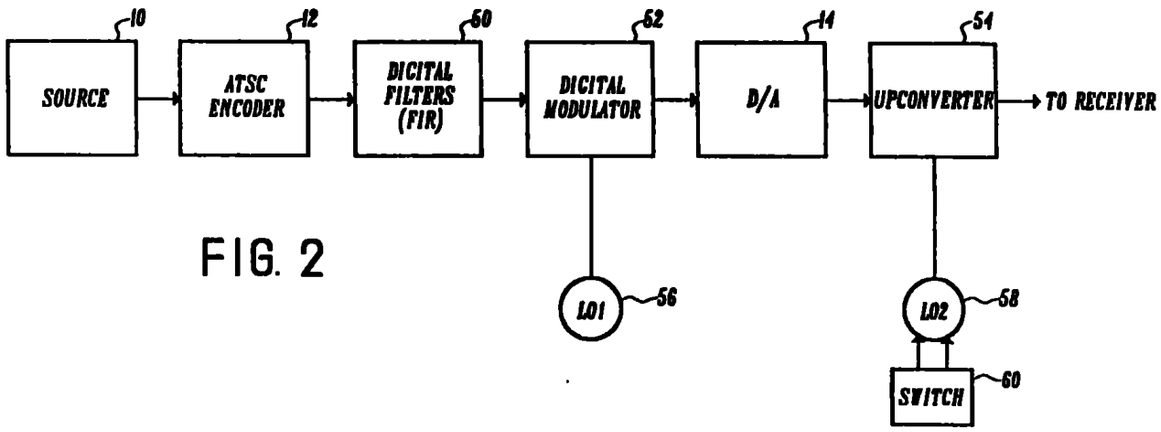


FIG. 2

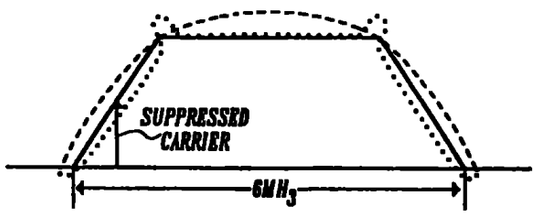


FIG. 3

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224

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US99/00130

223

A. CLASSIFICATION OF SUBJECT MATTER
IPC(6) :H04N 5/21, 7/04, 5/44
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According to International Patent Classification (IPC) or to both national classification and IPC

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Minimum documentation searched (classification system followed by classification symbols)

U.S. : 348/607, 608, 720, 469, 470, 625, 627, 723, 724, 725, 726, 426

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US 5,559,561 A (WEI) 24 September 1996, col. 2 lines 54-67, col. 3 lines 46-67, col. 4 lines 1-27.	1-20
A	US 5,648,822 A (HULYALKAR) 15 July 1997, col. 5 lines 51-67, col. 6 lines 1-52.	1-20
A	US 5,602,602 A (HULYALKAR) 11 February 1997, col. 4 lines 19-67, Fig. 1, Fig. 2.	1-20
A	US 5,572,249 A (GHOSH) 05 November 1996, Fig. 1, Fig. 2.	1-20
A	US 5,686,966 A (DE LA CIERVA, SR.) 11 November 1997, Fig. 1A, Fig. 1B.	1-20

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

*	Special categories of cited documents:	"T"	later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"A"	document defining the general state of the art which is not considered to be of particular relevance	"X"	document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
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"O"	document referring to an oral disclosure, use, exhibition or other means		
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Date of the actual completion of the international search	Date of mailing of the international search report
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Name and mailing address of the ISA/US Commissioner of Patents and Trademarks Box PCT Washington, D.C. 20231 Facsimile No. (703) 305-3230	Authorized officer JEAN W. DESIR Jeni Hill Telephone No. (703) 308-9571

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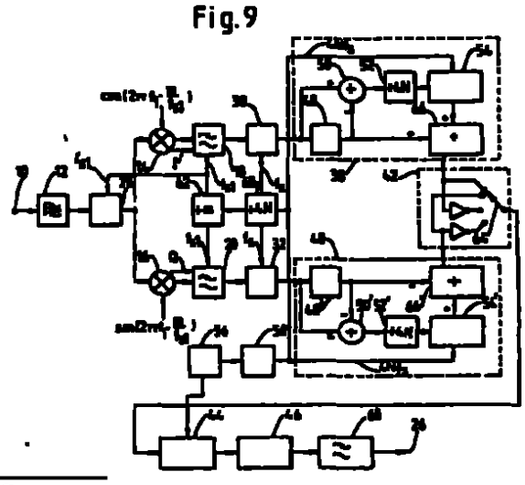
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(54) An arrangement for generating an SSB signal.

(57) An SSB generator in which a digitised analogue signal is processed in accordance with Weaver's method. The baseband I and Q signals are interpolated at a sampling frequency f_s in respective first interpolators (30, 32). The interpolated signals are frequency up-converted to an output SSB carrier frequency Nf_s by applying said signals to respective second interpolators (38, 40) in which the signals are sampled at $4Nf_s$. The samples at the outputs of the second interpolators (38, 40) are applied to a multiplexer (42) which provides a digitised version of the SSB signal which is converted to an analogue signal and filtered.

Fig.9



EP 0 323 675 A1

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AN ARRANGEMENT FOR GENERATING AN SSB SIGNAL

The present invention relates to an arrangement for generating an SSB signal and has applications in both the analogue and conventional digital implementations single sideband modulators constructed in accordance with what is termed Weaver's method. Such modulators are used in SSB generation in speech and data transmission for mobile radios.

Weaver's method is disclosed in Proceedings of the IRE "A Third Method of Generation and Detection of Single-Sideband Signals" Donald K. Weaver, Jr., pages 1703 to 1705. Figure 1 of the accompanying drawings illustrates a modulator made in accordance with Weaver's method. An analogue input signal at an input 10 is band limited in a bandpass filter 12 and then is applied to balanced mixers 14, 16 where it is mixed with quadrature related carrier signals cos ($\omega_0 t$) and sin ($\omega_0 t$), where $\omega_0 t$ is a carrier frequency corresponding to the centre of the passband of the signal from the filter 12. As a result of these mixing operations quadrature related signals I and Q are produced which have single sideband components at base band and unwanted components at high frequencies. The wanted components of the I and Q signals are selected by low pass filters 18, 20 and then frequency up-converted to their carrier frequency ($\omega_c t$) in balanced mixers 22, 24. The signals from these mixers are combined in a summing circuit 25 to provide an output on a terminal 26.

In data transmission a similar arrangement to that shown in Figure 1 is used to frequency up-convert the baseband I and Q modulated signals corresponding to the data. Weaver's method may also be used to generate constant envelope signals suitable for mobile radio digital transmission.

A drawback to Weaver's method is that it is difficult to implement the frequency up-converting or RF section as an integrated circuit. Reasons for this difficulty include the fact that matching of the mixers 22, 24 (Figure 1) is complicated and it may be necessary to run the mixers at a high power level to achieve the desired performance. Also it is necessary to make a pair of quadarature local oscillators capable of generating a wide range of frequencies. Any mismatch in the mixers or any imperfection in the quadrature RF signal generation may cause the unwanted signal levels to increase.

Mismatch problems mentioned above can be overcome by generating digital samples of the RF local oscillators and using digital multipliers instead of the RF mixers. This is shown in Figure 2 of the accompanying drawings which illustrate a conventional digital implementation of Weaver's method. In Figure 2 the output of the bandpass filter 12 is digitised in an analogue-to-digital converter 28 which is sampled at a clock frequency f_{s1}. The digitised signals are applied to mixers 14, 16 which respectively receive digitised local oscillator signals cos $[(2\pi f_1 . (m/f_{s1})]$ and sin $(2\pi f_1 . m/f_{s1})$. The I and Q signals from the mixers are low pass filtered in digital filters 18, 20 which are clocked at the frequency f_{s1} which satisfies the Nyquist sampling rate. The output signals which have been sampled at f_{s1} are applied to respective interpolators 30, 32 operating at a sampling frequency Nf_s. The outputs of the interpolators 30, 32 are frequency up-converted in digital multipliers 34, 36 to which are supplied quadrature carrier signals cos $[2\pi f_c . (n/Nf_s)]$ and sin $[2\pi f_c . (n/Nf_s)]$, respectively. The outputs from the digital multipliers 34, 36 are combined in the summing circuit 25 and the R.F. output appears on the terminal 26. This digital implementation of Weaver's method enables a performance improvement to be achieved at the expense of increased power consumption as digital multipliers require more power than the equivalent analogue mixers. The generation of RF quadrature signal samples raises another problem for radio transmitter integration when constant frequency sampling is required at RF. The sine and cosine samples can be generated by using either look-up tables or online processing but neither method is attractive for integration.

An object of the present invention is to provide an integratable SSB generator which is economical on its use of power.

According to a first aspect of the present invention there is provided an SSB generator comprising means for providing quadrature related digitised first and second input signals at baseband frequencies, interpolating means for frequency up-converting said first and second signals, and multiplexing means coupled to the interpolating means for multiplexing the frequency up-converted first and second input signals to provide a digitised SSB signal having a desired carrier frequency.

The interpolating means may comprise first and second interpolating means, the first interpolating means serving to sample respectively the first and second signals at a frequency (f_s) greater than the digitising frequency and the second interpolating means serving to sample the signals from the first interpolating means at a multiple of the first mentioned sampling frequency. The multiplexing means may be arranged to multiplex the samples at the sampling frequency of the second interpolating means.

Implementing the interpolating means as first

BNSDOCID: <EP 0323675A1 I >

and second interpolating means enables each to be adapted to the task required of it. The first interpolating means operates as a digital filter and serves to increase the repetition rate of the signal from a digital low pass filter to make it suitable for linear interpolation in the second interpolating means. The second interpolating means operates as a linear interpolator which lifts the frequency up to RF.

In an embodiment of the present invention the sampling frequency used in the second interpolation operation is 4N times the sampling frequency (f_s) used in the first interpolation operation, that is $4Nf_s$, and the multiplexing operation is carried out at a clock frequency of $4Nf_s$ to provide the SSB signal at Nf_s.

The multiplexed output may be interpolated for a third time in order to shift the sampling frequency $4Nf_s$ to a higher frequency outside the bandwidth of a low pass filter connected to a digital to analogue converter coupled to the multiplexing means.

The second interpolation means may be adapted to calculate the values of all the samples of the frequency up-converted first and second digitised signals. Alternatively power can be saved by only calculating the values of those samples which after multiplexing will contribute to the final signal. In order to do this only the odd-numbered samples in one of the first and second interpolated signals are calculated and only the even-numbered samples in the other of the first and second interpolated signals are calculated.

The present invention will now be explained and described, by way of example, with reference to the accompanying drawings, wherein:

Figure 1 is a block schematic diagram of an analogue modulator constructed in accordance with Weaver's method,

Figure 2 is a block schematic diagram of a digitised modulator constructed in accordance with Weaver's method,

Figure 3 is a block schematic diagram of an embodiment of an SSB modulator made in accordance with the present invention,

Figure 4 shows the impulse response of an interpolating filter,

Figure 5 shows the frequency response of an interpolating filter,

Figure 6 shows the impulse response of an interpolating filter,

Figure 7 shows two samples taken from the impulse response shown in Figure 6, which samples are separated in time by $1/f_s$,

Figure 8 is a block schematic diagram of a linear interpolation filter,

Figure 9 is a block schematic diagram of an embodiment of a SSB modulator including the linear interpolation filter shown in Figure 8, and

Figure 10 is a block schematic diagram of another embodiment of an SSB modulator including an alternative linear interpolation filter to that shown in Figures 8 and 9.

In the drawings, the same reference numerals have been used to indicate corresponding features.

Referring to Figure 3, the SSB modulator shown is essentially the same as that shown in Figure 2 as far as the interpolators 30, 32 which are operating at a sampling frequency, f_s, which is of the order of 30 times higher than the frequency at the top end of the bandwidth of the low pass filters 18, 20. The outputs from the interpolators 30, 32 are applied to second interpolators 38, 40 operating at a higher sampling frequency, Nf_s, than f_s. The outputs from the second interpolators 38, 40 are multiplexed in multiplexer 42 to provide an output comprising the SSB signal centred on $Nf_s/4$ plus the sampling frequency component in the output from the multiplexer, the output is applied to a third interpolator 44 which has a sampling frequency which is an integer multiple of Nfs. A digital to analogue converter 46 is connected to the third interpolator 44 and provides an output on the terminal 28. By this technique the frequency band of the SSB signal is maintained around f_s whilst the Nf_s component is frequency up-converted to a frequency which lies outside the pass bandwidth of a filter in the DAC 46. If desired a superhet stage may be connected to the output 28 to increase the frequency of the SSB signal which in such a situation represents an IF signal.

Figure 4 shows the impulse response of an interpolating filter which may comprise the second interpolator 38 or 40. The ordinate represents amplitude and the abscissa time. The filter's response lasts two bit periods $\pm T_s$, where $T_s = 1/f_s$, and has a $\sin^2 x/x^2$ type of transfer function. The response shape is sampled at N times the sampling frequency f_s, that is at Nf_s. As shown in Figure 5, the frequency response of the interpolating filter has zero crossings at each integer multiple of f_s, but the signal at Nf_s will not be attenuated.

Thus when the signal samples, sampled at f_s, are applied to the interpolating filter, the filter as a consequence will eliminate all the signals between f_s and Nf_s. The output produced comprises an interpolated version of the input signal having a new sampling frequency of Nf_s.

A simplified way of realising an interpolating filter will now be described with reference to Figures 6, 7 and 8. The value of N is determined by the highest required integer number used to increase the sampling frequency f_s. When using the interpolating filter having an impulse response as shown in Figure 6, the filter response length is $-T_s$ to T_s which equals $2/f_s$. Furthermore at any given

BNSDOCID <EP 0323675A1 I >

time there will be only two samples x_1 and x_2 in the interpolating filter which will appear at instances separated by $T_s = 1/f_s$. Therefore in realising the interpolating filter as a transversal filter only two multipliers at a time are required together with an addition circuit to sum the multiplied signal samples. The resultant signal, y, appearing at the output of the filter will be given by:

$y = ax_1 + bx_2$ (1)

where a and b are the coefficients of the convolution of the filter at the nth and (n+N)th taps. Since the impulse response of the filter as shown in Figure 8 has a triangular shape, the coefficients of a and b are given by:

$a = n/N$ (2)

$b = 1-n/N$ (3)

where N is the integer multiple used to increase the sampling frequency.

Substituting equations (2) and (3) into equation (1) leads to:

$y = ax_1 + bx_2 = x_2 + n(x_1 - x_2)/N.$

This means that if the sampling frequency f_s is to be increased by a factor of N, then one obtains the difference between adjacent samples, x_1 and x_2, at the sampling frequency f_s and divide the result by the interpolation factor, N. The resulting quotient is then taken as an increment and added to the sample x_2 at every instant n to produce the interpolated samples of the input signal at the new sampling frequency Nf_s.

Figure 8 illustrates an embodiment of the incrementing filter realised to implement

$y = x_2 + n(x_1 - x_2)/N.$

The signal sampled at f_s is applied to a shift-register or tapped delay line 46. Samples x_1 and x_2 separated by $1/f_s$ seconds are tapped off and applied to a subtractor 50 which produces the difference or increment (x_1-x_2) which is applied to a divider 52 in which the increment is divided by N. The quotient from the divider is added to the sample x_2 in an accumulator 54 to produce the interpolated samples of the input signal at the new sampling frequency Nf_s.

Figure 9 is a block schematic diagram of an SSB modulator made in accordance with the present invention. The front end portion corresponds to that described with reference to and as shown in Figure 3 and in the interests of brevity will not be described again. The second interpolators 38, 40 are based on the circuit as shown in Figure 8 with some modifications. In the case of the second interpolator 40, the circuit stages which correspond to those used in the second interpolator 38 have been referenced using primed reference numerals, that is 48', 50', 52' and 54'. However the differences are divided by 4N in the dividers 52 and 52' and the accumulators 54, 54' are clocked at $4Nf_s$. The effect of doing this is that the increment is obtained by dividing the difference by 4N. The increment is then accumulated on the delayed signal sample x_2 at regular clocking points. The accumulation starts every time a new sample is loaded into the shift register or tapped delay line 48, 48' and continues for 4N clock periods until a new sample is loaded into the shift register so that the process is repeated for a new value of x_2.

The clock signal $4Nf_s$ is derived from a master clock 58 operating at a higher frequency than $4Nf_s$ and divided down in a divider 58. This divided down signal at $4Nf_s$ is divided by 4N in another divider 60 to provide clock signals at f_s, where f_s has a value of the order of 100KHz, to the interpolators 30, 32. The clock signal f_{s1} for the A to D converter 28 and the digital low pass filters 18, 20 is derived by dividing f_s by m, where m is of the order of 8, in a divider 62.

The multiplexer 42 provides an SSB signal at a nominal carrier frequency Nf_s sampled at $4Nf_s$. The multiplexer 42 comprises the electronic equivalent of four pole switch 64 operating at $4Nf_s$. Proceeding from top to bottom in Figure 9, the first and second poles are respectively connected to addition stages 66, 66' of the second interpolators 38, 40 which stages provide digital values of the samples. The third and fourth poles provide inverted forms of the signals on the first and second poles, respectively. The output from the multiplexer 42 comprises a digitised form of the SSB signal at the nominal carrier frequency Nf_s and the unwanted sampling frequency at $4Nf_s$. In order to eliminate the unwanted sampling frequency the multiplexer output is applied to the third interpolator 44 in which it is sampled at the clock frequency of the master clock 58, which frequency is above the bandwidth of a low pass filter 68 connected to the output of the D to A converter 46. By this operation the sampling frequency $4Nf_s$ is eliminated by interpolation so that it is outside the bandwidth of the low pass filter 68.

The operation of the circuit shown in Figure 9 will now be summarised. In order to produce the desired SSB signal, the required quadrature oscillator samples will be 1, 0, -1, 0, 1, 0 and 0, 1, 0, -1, 0,1 in the I and Q channels, respectively. In order to realise the RF section, these I and Q signal samples appearing at the outputs of the base band filters 18, 20 are interpolated to a sampling frequency $4Nf_s$ in the interpolators 30, 38 and 32, 40. These samples are multiplied with the RF oscillator samples (see above) so that the alternate products of the multiplication in the I and Q channels become equal to zero. The multiplexer 42 selects samples alternately from the I and Q channels, which samples after D to A conversion and filtering form the SSB signal on the output 26.

In order for the SSB modulator to be able to

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operate at a range of carrier frequencies it is necessary to vary the sampling frequencies used. In order to do this then for a certain carrier frequency f_c, the SSB signal sampling frequency $4Nf_s$ is selected to be 4 times higher than f_c. The sampling frequency f_s for the interpolators 30, 32 is chosen by dividing $4Nf_s$ by 4N in the stage 60. The frequency f_{s1} is produced by dividing f_s by m in the stage 62. The master clock frequency is generated using a phase locked loop. The local oscillator frequency f_1 is kept constant.

Two numerical examples of how the frequencies are calculated will now be given. It will be assumed that it is desired to cover SSB generation over the frequency range 0 to 10 MHz and in consequence the bandwidth of the low pass filter 68 is 10 MHz.

In the first numerical example it will be assumed that the SSB signal is to be transmitted at a frequency

Nf_s = 5MHz

$4Nf_s$ = 20MHz

f_s should be of the order 100kHz and therefore

N = 5000/100 = 50 (and 4N = 200).

If m = 8 then f_{s1} = 12.5KHz.

As $4Nf_s$ = 20 MHz is outside the bandwidth of the low pass filter 68 it will be attenuated to an acceptable level. In this situation there is no need to use the third interpolator 44.

In the second numerical example it will be assumed that the SSB is to be transmitted at a frequency

Nf_s = 1MHz

so that $4Nf_s$ = 4MHz, f_s is maintained at 100 kHz and therefore

N = 1000/100 = 10 (and 4N = 40). The values of m and f_{s1} are maintained at 8 and 12.5kHz, respectively.

As $4Nf_s$ = 4 MHz which is within the 10 MHz bandwidth of the low pass filter 68, the interpolation factor and the sampling frequency of the third interpolator 44 are set to 3 and 12 MHz, respectively. Thus the divider 58 divides the 12 MHz by 3 to obtain the desired $4Nf_s$.

If a new carrier frequency is chosen then the calculations are repeated and the master clock frequency and the respective divisors are determined accordingly.

By operating the multiplexer 42 at $4Nf_s$ odd numbered samples are selected from say the I channel and even numbered samples are selected from the other, Q channel. As a result the non-selected samples are not used. Consequently there is no need to calculate the even numbered samples in the I channel and the odd numbered samples in the Q channel. Thus the interpolation equation given in the description of Figure 8 can be modified to form equations for y_I and y_Q where

$y_I = x_2^I + 2n\,(x_1^I - x_2^I)/4N$

and $y_Q = X_2^Q + (2n+1)\,(x_1^Q - x_2^Q 4N$.

As a result of this modification the second I and Q interpolators 38, 40 are operated at $2f_s$ and the signals are multiplexed at $4Nf_s$.

An embodiment of the invention in which the modified equations are implemented is shown in Figure 10. In the interests of brevity only the differences between the embodiments of Figures 9 and 10 will be discussed. The master clock 56 is connected to the divider 58 which produces an output signal at $2NF_s$. The multiplexer 42 is still clocked at $4Nf_s$. The signal at $2Nf_s$ is supplied firstly to the divider 60 where it is divided down by 2N to provide the clock signal f_s to the interpolators 30, 32 and secondly to increment a counter 70 two steps at a time. The counter 70 is reset at the instants when data is loaded into the shift registers 48, 48'. The output of the counter 70 is applied to a multiplier 54' and to an adder 74 in which it is added to +1 provided by a stage 72. The sum 2n+1 is supplied to the multiplier 54. In the multiplier 54 the number obtained from the divider is multiplied by 2n+1 to produce the increments for the I channel. Similarly in the multiplier 54' the output of the divider 52' is multiplied by 2n to produce the increments for the Q channel.

The SSB modulators shown in Figures 9 and 10 are capable of being fabricated as integrated circuits. If desired a signal processor may be used to realise the base band mixers 14, 16 and the interpolators 30,32, 38 and 40.

Claims

1. An SSB generator comprising means for providing quadrature related digitised first and second input signals at baseband frequencies, interpolating means for frequency up-converting said first and second signals, and multiplexing means coupled to the interpolating means for multiplexing the frequency up-converted first and second input signals to provide a digitised SSB signal having a desired carrier frequency.

2. An SSB generator as claimed in Claim 1, characterised in that the interpolating means comprises first and second interpolating means, the first interpolating means serving to sample respectively the first and second signals at a frequency (f_s) greater than the digitising frequency and the second interpolating means serving to sample the signals from the first interpolating means at a multiple of the first mentioned sampling frequency, and in that the multiplexing means is arranged to multiplex the samples at the sampling frequency of the second interpolating means.

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3. An SSB generator as claimed in Claim 2, characterised in that the second interpolating means samples the signals at the output of the first interpolating means at a frequency 4N times the first mentioned sampling frequency, that is at $4Nf_s$, and in that the multiplexing means is operated at 4Nfs to provide an output signal at Nf_s.

4. An SSB generator as claimed in Claim 2 or 3, characterised in that a digital to analogue converter (DAC) is coupled to the multiplexing means and a low pass filter connected to the output of the DAC.

5. An SSB generator as claimed in Claim 4, characterised in that a third interpolating means is connected to an output of the multiplexing means, the third interpolating means being operated at a sampling frequency greater than the sampling frequency of the second interpolating means in order to place the interpolated latter sampling frequency outside the bandwidth of the low pass filter.

6. An SSB generator as claimed in any one of Claims 2 to 5, characterised in that the second interpolating means comprises means for implementing the equation

$y = ax_1 + bx_2 = x_2 + n(x_1 - x_2)/N$ where x_1 and x_2 are samples separated in time by $1/f_s$ of the impulse response of a filter, $(x_1 - x_2)$ is the difference in amplitude of the two samples and N is the multiple of the sampling frequency (f_s) of the first interpolating means to obtain the SSB signal.

7. An SSB generator as claimed in any one of Claims 2 to 5, characterised in that the second interpolating means is arranged to operate on odd numbered samples of one of the quadrature related signals received from the first interpolating means and on even numbered samples of the other of the quadrature related signals received from the first interpolating means.

8. An SSB generator as claimed in Claim 7, characterised in that the second interpolating means comprises means for implementing the interpolation equations

$y_I = X_2^I + 2n(x_1^I - x_2^I)/4N$

and $y_Q = x_2^Q + (2n+1)(x_1^Q - x_2^Q)/4N$ where y_I and y_Q are the quadrature related signals produced by the second interpolating means, x_1^I, x_2^I and x_1^Q, X_2^Q are samples separated in time by $1/f_s$ of the impulse response of the filter, $(x_1^I - x_2^I)$ and $(x_1^Q - x_2^Q)$ are the differences in amplitude between the respective pairs of samples and N is the multiple of the sampling frequency (f_s) of the first interpolating means to obtain the SSB signal.

9. An SSB generator as claimed in Claim 6, 7 or 8, characterised in that the second interpolating means comprises, for each of the quadrature related signals, shift register means, means for sampling the amplitude of the bits stored, differencing means for obtaining the difference between the amplitudes of two bits separated in time by $1/f_s$, and means for adding the difference obtained to the amplitude of the second sample.

10. An SSB generator as claimed in any one of Claims 2 to 9, characterised in that the multiplexing means comprises means for providing the samples at the higher sampling frequency of the respective quadrature related signals and the inverted values of said respective quadrature related signals and means for scanning said sample and said inverted values.

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Fig.1

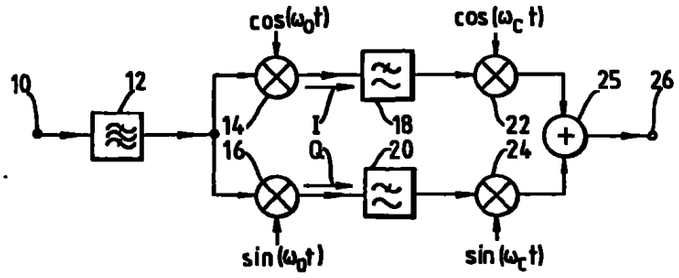
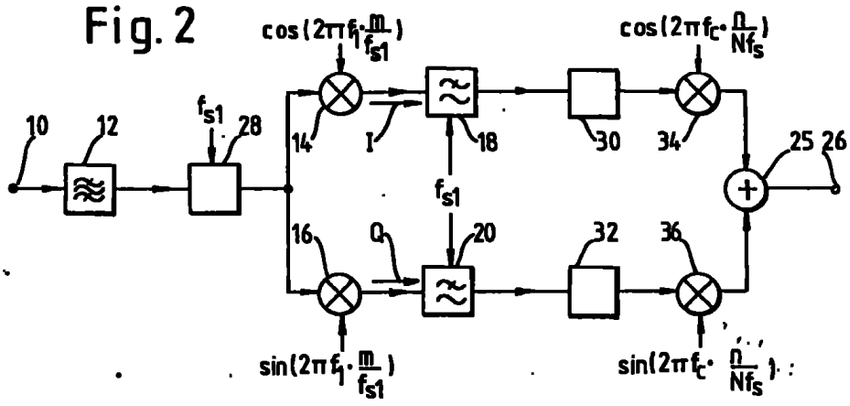


Fig. 2



EP 0 323 675 A1

Fig. 3

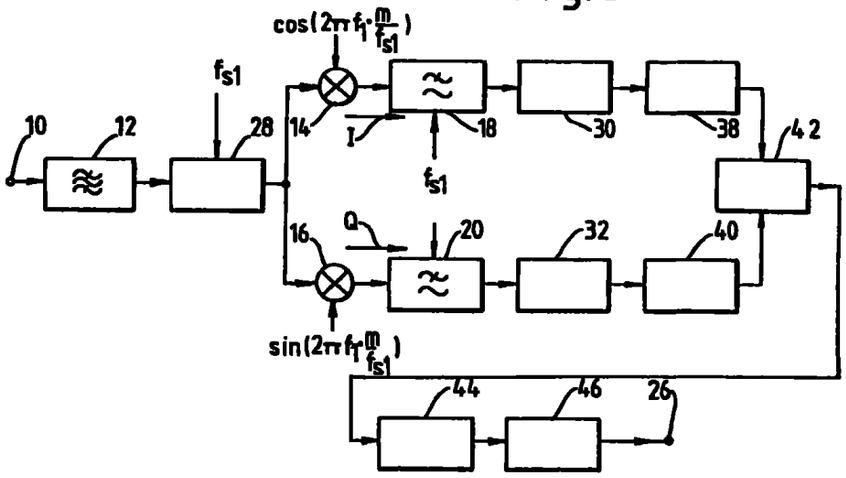
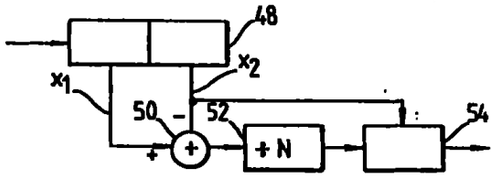

cos(2πf1·m/fs1)
fs1
10
12
28
14
I
18
30
38
42
16
Q
20
32
40
sin(2πf1·m/fs1)
44
46
26

Fig. 8


48
x2
x1
50
52
54
+N

27 12 88

229

Fig.4

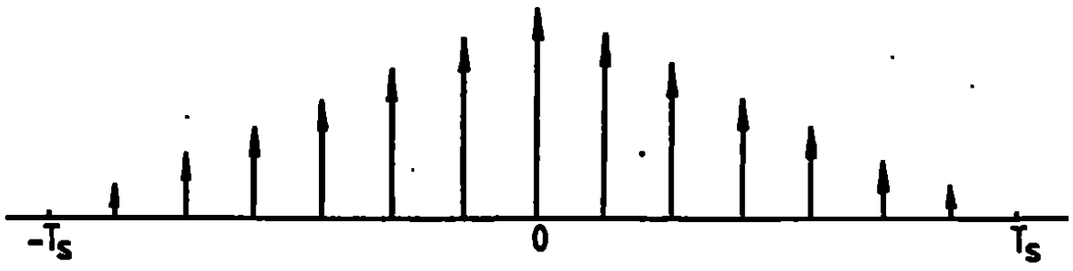


Fig.5

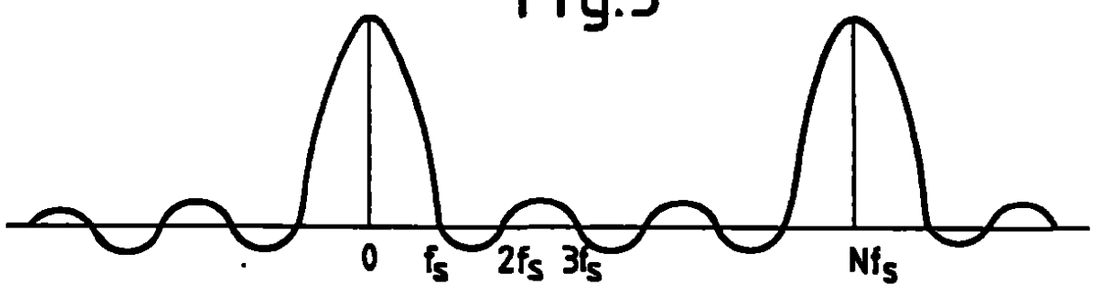
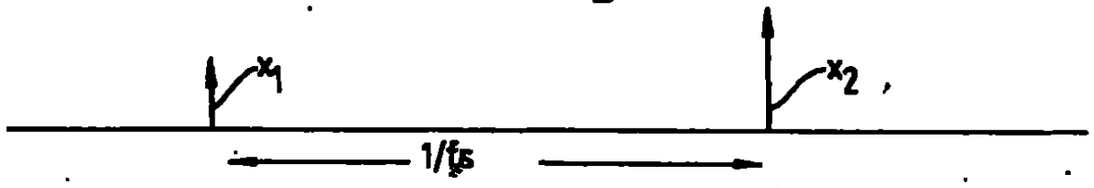


Fig.6



Fig. 7



3-V-PHB 33431

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Fig.9

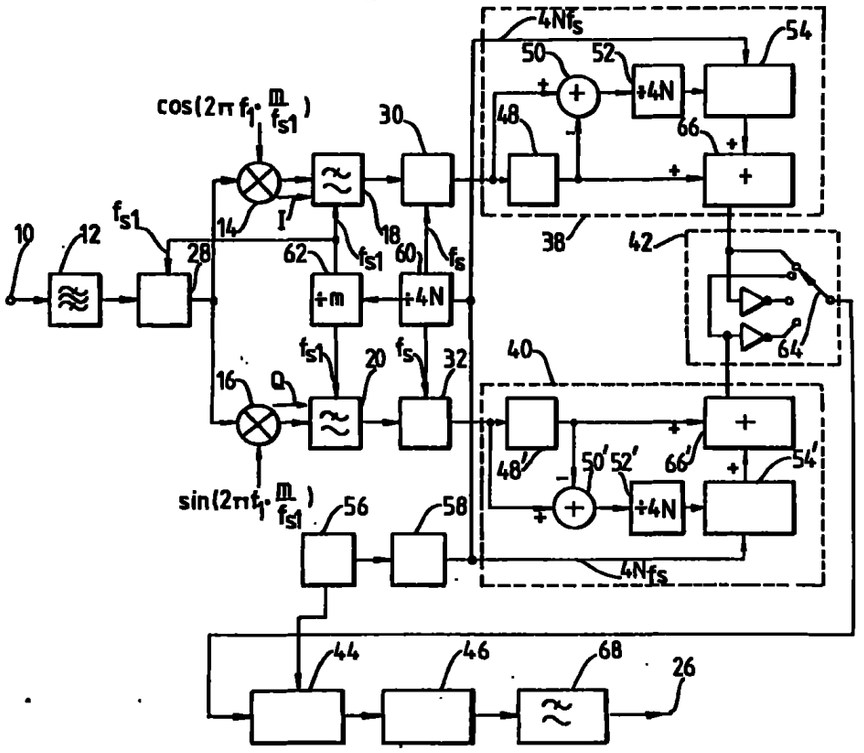
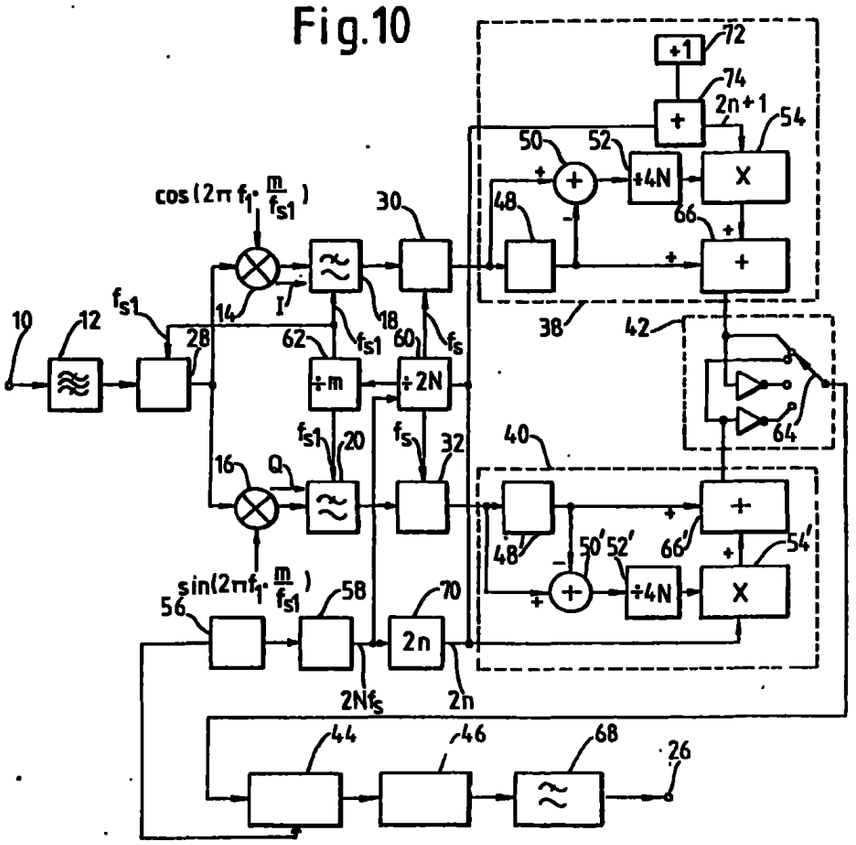

4Nfs
50
52
+4N
54
cos(2π f1 m/fs1)
30
48
66
10
12
fs1
28
14
I
18
62
fs1
60
fs
38
42
+m
+4N
64
fs1
20
fs
32
40
16
Q
48'
50'
52'
66'
54'
sin(2π f1 m/fs1)
56
58
4Nfs
44
46
68
26

Fig.10



EP 0 323 675 A1

European Patent Office

EUROPEAN SEARCH REPORT

Application Number

EP 88 20 3015

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int. Cl.4)
A	US-A-4 086 536 (ACKER) * Figure 11; column 8, lines 7-35; column 10, lines 51-58, column 13, lines 65-69; column 17, lines 23-26; column 18, line 4 - column 20, line 27 * ---	1-10	H 03 C 1/60
A	US-A-3 997 773 (VAN ESSEN et al.) * Figure 1; abstract * ---	1-3,5-9	
A	IEEE TRANSACTIONS ON COMMUNICATIONS, vol. COM-30, no. 7, part 1, July 1982, pages 1629-1641, New York, US; A.G. CONSTANTINIDES et al.: "An efficient and modular transmultiplexer design" * Figure 3; page 1631, lines 21-24 * -----	1-3,5-9	

TECHNICAL FIELDS SEARCHED (Int. Cl.4)

H 03 C
H 03 D
H 03 H
H 04 B
H 04 L

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	08-03-1989	PEETERS M.M.G.

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
Y : particularly relevant if combined with another document of the same category
A : technological background
O : non-written disclosure
P : intermediate document

T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

EPO FORM 1503 03.82 (P0401)

231

STATE OF ILLINOIS

SECRETARY OF STATE

SPRINGFIELD, ILLINOIS

APOSTILLE
APOSTILLE

(Convention de La Haye du 5 Octobre 1961)

1. Country: United States of America

   This public document

2. has been signed by DONNA B MORGAN

3. acting in the capacity of NOTARY PUBLIC, WILL COUNTY

4. bears the seal/stamp of STATE OF ILLINOIS

Certified

5. Chicago, Illinois

6. JANUARY 27, 2003

7. by the Secretary of State, State of Illinois

8. No. 7833

9. Seal/Stamp:

10. Signature:

NOTARIA 16
03 FEB 2003
FOTOCOPIA CONCUERDA EXACTAMENTE CON EL ORIGINAL QUE TUVE A LA VISTA

Jesse White

JESSE WHITE
SECRETARY OF STATE
STATE OF ILLINOIS

I-168.2

SCHIFF HARDIN & WAITE
A Partnership Including Professional Corporations

6600 Sears Tower, Chicago, Illinois 60606-6473
Telephone (312) 258-5500 Facsimile (312) 258-5700

Susan Collins (312) 258-4969

Chicago
Washington
New York
Merrillville
Dublin
Lake Forest

232

221

January 27, 2003

Secretary of State of Illinois
Index Department
17 N. State Street, Suite 1137
Chicago, IL 60602

I, Susan Collins, do state that this is a true copy of the original Assignment of Patent Application U.S. Serial No. 09/632/545 filed August 4, 2000 from Thomas P. Horwitz to Zenith Electronics Corporation for use in Colombia.

Date: January 27, 2003 Signature: Susan Collins
Name: Susan Collins
Title: Patent Paralegal

State of Illinois)
: ss
County of Cook)

Sworn to and scribed before me, a Notary Public, this 27th day of Jan., 2003.

Notary Public

OFFICIAL SEAL
DONNA J. MORGAN
NOTARY PUBLIC, STATE OF ILLINOIS
MY COMMISSION EXPIRES 2-26-2003

NOTARIA 16
03 FEB 2003
ESTA FOTOCOPIA COINCIDE EXACTAMENTE CON EL ORIGINAL QUE TUVE A LA VISTA

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233

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ASSIGNMENT

In Consideration of One Dollar and other good and valuable consideration, the receipt of which is hereby acknowledged, I:

THOMAS P. HORWITZ of ELGIN, ILLINOIS

sell and assign to ZENITH ELECTRONICS CORPORATION, a corporation of Delaware whose principal place of business is located at 1000 Milwaukee Avenue, Glenview, Illinois 60025, United States of America, its successors and assigns, all right, title, and interest in the United States and throughout the world in and to the improvements in "LOW COST VSB ENCODER AND RF MODULATOR FOR SUPPLYING A SUBSTANTIALLY 6 MHZ VSB SIGNAL TO DIGITAL TELEVISION RECEIVER" invented by me, and the application for United States Letters Patent therefor executed concurrently herewith, and any corresponding applications filed anywhere in the world, and all patents which may be granted therefor, and all divisions, reissues, continuations and extensions thereof, and authorize and request the Commissioner of Patents to issue all patents on said improvements or resulting therefrom to said Corporation as assignee of the entire interest, and covenant that I have full right so to do, and agree that I will communicate to said Corporation or its representatives any facts known to us respecting said improvements and testify in any legal proceedings, sign all lawful papers, execute all divisional, continuing and reissue applications, make all rightful oaths, and generally do everything possible to aid said Corporation, its successors, assigns and nominees, to obtain and enforce proper patent protection for said invention in all countries, and I authorize any of our principal attorneys of record in said application to insert on this assignment the filing date and serial number of said application, when known.

IN TESTIMONY WHEREOF, I have hereunto set my hand and seal this 3rd day of August, 2000.

_____ L.S.

STATE of Illinois
COUNTY of Cook

On this 3 day of August, 2000, before me, a Notary Public in and for the State and County aforesaid, personally appeared Thomas P. Horwitz to me personally known as the individual who executed the foregoing assignment, and who acknowledged to me that he executed the same of his own free will and for the purposes therein set forth.

_____

OFFICIAL SEAL
SHARON KROLL
NOTARY PUBLIC STATE OF ILLINOIS
MY COMMISSION EXP. NOV.19,2000

(SEAL)

The above assignment covers application Serial No. 09/633,345 filed August 4, 2000.

_____
Jack Kail
Registration No. 27,588

233
234

Traducción Oficial No. 2003/259 de un documento escrito en inglés el cual para su identificación se sella con el sello del traductor al tiempo con la presente. Esta traducción ha sido preparada por Hilda Aguilera de Piedrahita, traductora oficial debidamente reconocida por el Ministerio de Relaciones Exteriores y autorizada por el Ministerio de Justicia y del Derecho por Resolución No. 323 de 1980.

CESION

En consideración de Un Dólar y otras consideraciones buenas y valiosas, el recibo de las cuales por el presente se reconoce, Yo:

THOMAS P. HORWITZ de ELGIN, ILLINOIS

vendo y cedo a ZENITH ELECTRONICS CORPORATION, una sociedad de Delaware con asiento principal de negocios localizado en 1000 Milwaukee Avenue, Glenview, Illinois 60025, Estados Unidos de América, sus sucesores y causahabientes, todos los derechos títulos y participación en los Estados Unidos y en todo el mundo, en y para las mejoras en "CODIFICADOR VSB Y MODULADOR RF DE BAJO COSTO PARA SUMINISTRAR UNA SEÑAL VSB SUSTANCIALMENTE DE 6 MHZ A UN RECEPTOR DE TELEVISION DIGITAL" inventado por mí, y la solicitud para Certificado de Patentes en los Estados Unidos para la misma suscrita concurrentemente con la presente, y cualesquiera solicitudes correspondientes presentadas en cualquier parte del mundo, todas las patentes que puedan ser otorgadas para la misma, y todas las divisiones, reexpediciones, continuaciones y extensiones de la misma,, y autorizamos y

NOTARIA 6 ... 05 FEB 2003 ...

TRADUCCION OFICIAL No. 2003/259
HILDA AGUILERA DE PIEDRAHITA
INTERPRETE OFICIAL

solicitamos al Comisionado de Patentes para expedir todas las Patentes para dichas mejoras o las que resultaren de las mismas a dicha sociedad como Cesionaria de la participación entera, y declaro que tengo pleno derecho para hacerlo, y acuerdo que comunicaré a dicha sociedad o a sus representantes cualesquiera hechos conocidos por nosotros con respecto a dichas mejoras y a testificar en cualquier proceso legal, firmar todos los papeles legales,, suscribir todas las solicitudes de división, continuación y reexpedición,, efectuar todos los juramentos de acuerdo con la ley, y en general hacer todo lo posible para ayudar a dicha sociedad, sus sucesores,. causahabientes y personas nombradas para obtener y hacer cumplir la protección propia de la patente para dicha invención en todos los países, y autorizo a cualquiera de nuestros apoderados principales registrados en dicha solicitud para insertar en esta cesión la fecha de presentación y número de serie de dicha solicitud, cuando se conozca.

EN TESTIMONIO DE LO CUAL, he puesto mi firma y sello este día 3 de Agosto de 2000.

(Firmado: Ilegible)

ESTADO de Illinois

CONDADO de Cook



TRADUCCION OFICIAL No. 2003/259
HILDA AGUILERA DE PIETRANTONIO
INTERPRETE OFICIAL

En este día 3 de Agosto de 2000,, ante mí, un Notario Público en y para el Condado anteriormente nombrado, compareció personalmente Thomas P. Horwitz, a quien conozco personalmente como el individuo que suscribió la Cesión precedente, y quien reconoció ante mí que él suscribió la misma como su acto propio de voluntad libre y para los propósitos allí señalados.

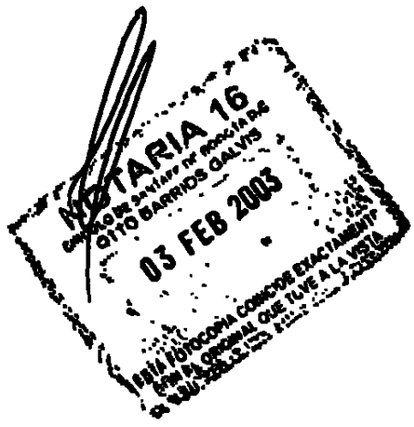
(Firmado: Sharon Knoll)

Sello a tinta: Sello Oficial, Sharon Knoll, Notario Público para el Estado de Illinois, Mi comisión expira el 19 de Noviembre de 2000.
(Sello)

La anterior cesión cubre la Solicitud con No. De Serie 09/632.545 presentada el 4 de Agosto de 2000.

(Firmado: Ilegible)
Jack Kail
Registro No. 27.566

SCHIFF HARDIN & WAITE
Una sociedad que incluye Sociedades profesionales
6600 Sears Tower, Chicago, Illinois 60606-6473
Teléfono: (312) 258-5500 Fax (312) 258-5700



236
237

Chicago, Washington, Nueva York, Marrillville, Dublin, Lake Forest.

Susan Collins (312) 258-4969

Enero 27,2003.

Secretario de Estado de Illinois

Departamento de Indices

17N. State Street, Suite 1137

Chicago, Illinois 60602

Y , Susan Collins, declaro que esta es una copia verdadera del original de la cesión de la Solicitud de Patente con No. De Serie U.S. 09/632/545 presentada el 4 de Agosto de 2000 de Thomas P. Horwitz a Zenith Electronics Corporation para uso en Colombia.

Fecha: Enero 27, 2003.

Firma: (Firmado)

Nombre: Susan Collins

Cargo: Paralegal en Patentes



Estado de Illinois)

: ss

Condado de Cook)

2003/259

238

Juramentado y suscrito ante mí, un Notario Público, este día 27 de Enero de 2003.

(Firmado: Ilegible)

Notario Público

Sello a tinta: Sello Oficial , Donna B. Morgan, Notario Público del Estado de Illinois,

Mi comisión expira el 2-25-2003.

ESTADO DE ILLINOIS

SECRETARIA DE ESTADO

SPRINGFIELD, ILLINOIS

APOSTILLA

(Convención de la Haya de Octubre 5 de 1961)

1. País: Estados Unidos de América

   Este documento público

2. Ha sido firmado por DONNA B. MORGAN

3. Actuando en calidad de Notario Público, Condado de Will

4. Lleva el sello/estampilla del Estado de Illinois

CERTIFICADO

5. En Chicago, Illinois

6. Enero 27 de 2003.

NOTARIA 16 03 FEB 2003 ESTA FOTOCOPIA CONCUERDA EXACTAMENTE CON EL ORIGINAL QUE TUVE A LA VISTA

20031259

238

239

7. Por El Secretario de Estado , Estado de Illinois

8. Número 7833

9. Sello (Sello Redondo en Relieve)

10. Firma (Firmado)

JESSE WHITE

Secretario de Estado

Estado de Illinois

ES TRADUCCIÓN FIEL Y COMPLETA de la cual se deja copia en los archivos de esta oficina para su confrontación y a la cual me remito.

Bogotá, Enero 30 de 2003.

TRADUCCION OFICIAL No. 2003/259
SILDA AMURLEGA DE PEDRAHITA
INTERPRETE OFICIAL
[illegible]

NOTARIA 16
CIRCULO NOTARIAL DE BOGOTA D.C.
OTTO HARPHOY GALVIS
03 FEB 2003
ESTA FOTOCOPIA CONCUERDA EXACTAMENTE CON EL ORIGINAL QUE TUVE A LA VISTA

239

SUPERINTENDENCIA Y COMERCIO
Radicacion : 03007467 00000000 Folios: 225
Fecha (AMD): 2002-08-03 13:49:21
Tramite : 011 PCT-PATENT 279 FINE HACI 411 PRESENTE
Dependencia: 2000 DIVISION DE NUEVAS CREACIONES

240
239

REQUISITOS NECESARIOS PARA PRESENTACION Y AGILIZACION DEL TRAMITE

	USUARIO	RADICADOR
PATENTE DE INVENCIÓN MODELO DE UTILIDAD	( )	( )
- Indicación de que se solicita la concesión de una patente.		( )
- Datos de identificación del solicitante o de la persona que se presenta la solicitud.		( )
- Descripción de la invención.		( )
- Dibujos de ser estos pertinentes.		( )
- Comprobante de Pago de las tasas establecidas.		
COMPLETA ( ) INCOMPLETA	( )	( )
DISEÑO INDUSTRIAL ( )		
- Indicación de que se solicita el registro de Diseño Industrial	( )	( )
- Datos de identificación del solicitante o de la persona que presenta	( )	( )
- la solicitud.		
- Representación gráfica y fotográfica del Diseño Industrial o muestra del Material que incorpora el diseño.	( )	( )
Comprobante de pago de las tasas establecidas	( )	( )
COMPLETA ( ) INCOMPLETA	( )	( )
ESQUEMA DE TRAZADO ( )		
- Indicación de que solicita el registro de un Esquema de trazado.	( )	( )
- Datos de identificación del solicitante o de la persona que presenta La solicitud.	( )	( )
Representación gráfica del esquema de trazado	( )	( )
Comprobante de pago de las tasas establecidas.	( )	( )
COMPLETA ( ) INCOMPLETA	( )	( )

Firma Responsable: ANDREA PASSEU.

Fecha:



MINISTERIO DE RELACIONES EXTERIORES

006240

CUYA FIRMA APARECE EN EL DOCUMENTO DESEMPEÑA LAS FUNCIONES INDICADAS

NO SE ASUME RESPONSABILIDAD DEL TEXTO

2003 JAN 30 P 2:53

JEFE DE LEGALIZACIONES
SANTAFE DE BOGOTA D.C.

Industria y Comercio
SUPERINTENDENCIA

240
241

2020
Bogotá, D.C.,

Doctor:
Dario Cardenas Navas
Apoderado
Zenith Electronics Corporation

Oficio N° 07248

ASUNTO:		
	Radicación N°	03-7487
	Solicitud internacional	PCT/US01/23211
	Fecha inicio fase nacional	04/03/2003
	Trámite	011
	Evento	379
	Actuación	430
	Folios	001

Como resultado del examen de forma, realizado con fundamento el el artículo 38 de la Decisión 486 de la Comisión de la Comunidad Andina, artículo 27 del Tratado de Cooperación en Materia de Patentes (PCT), regla 51 bis del Reglamento y Circular Única de la Superintendencia de Industria y Comercio, se le comunica que:

☐ Los dibujos adjuntos no son legibles o copiables para trámites posteriores de publicación y título.

En cumplimiento de lo dispuesto por el artículo 39 de la Decisión 486, se le requiere para que dentro del término de dos meses siguientes a la fecha de notificación del presente oficio, complete los requisitos anteriormente enunciados. En caso de no dar cumplimiento al presente requerimiento, la solicitud se considerará abandonada y perderá su prelación.

Notifíquese el presente oficio conforme a lo dispuesto en el numeral 5.2, literal e), del capítulo quinto del título primero de la Circular Única N° 10 de 2001.

ALIX CARMENZA CESPEDES DE VERGEL
Jefe de la División de Nuevas Creaciones.

05 SET. 2003

El anterior requerimiento fue notificado por fijación en lista de fecha_____.

202047