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Polyphase Modem for Frequency-Division Multiplex

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Polyphase Modem for Frequency-Division Multiplex

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1. Introduction

The large volume of telephone traffic now being carried by wide-band frequency-division-multiplex carrier circuits has stimulated a search for less-expensive translation equipment. If the cost of a multiplex system is related to the individual circuit functions, by far the greatest part is in the modulators and filters in the voice-to-group translation and *vice versa*.

The object of the present development was a system with a lower sensitivity to variations in component values than existing equipment and which could be translated into an integrated-circuit form and so take advantage of their reducing cost.

The polyphase modulator can more readily accept the limited accuracy and stability of integrated components because the required discrimination is provided at a low frequency and unwanted 1st-order modulation products fall either within the channel frequency band or at a frequency outside the group band.

2. Analysis

2.1 Operation

The basic system is shown in Figure 1, and for ease of explanation only one path will be considered. The voice-frequency signal to be translated enters *via* a band-limiting network and is operated on by modulator *M1*. The carrier signal f_1 is selected to be at the center of the input voice-frequency band. The output of *M1* then consists of upper and lower sidebands as shown in Figure 2(b). Because the carrier frequency is at the voice-frequency center, the lower sideband extends to zero frequency and is folded over on itself. The output of the modulator *M1* is applied to the low-pass filter *F1* which transmits only the folded difference sideband (Figure 2(c)). This signal is operated on by a second modulator *M1′* whose carrier frequency f_2 lies at the center of the desired output band. The upper and lower sidebands so produced represent superimposed, erect, and inverted copies of the original input band, translated to a new position in the frequency spectrum (Figure 2(d)).

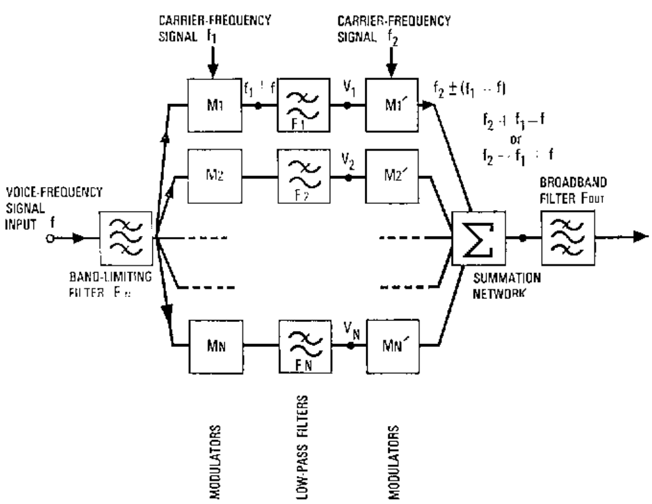

Figure 1 – Basic modulation system.

Now consider Figure 1 again, Here there are N identical paths with modulators *M1* through *MN* at the beginning of each path. Each modulator is supplied with the voice-frequency signal f and a carrier-frequency signal f_1. The latter signal is at a different phase for each modulator. For example, the carrier at *M2* lags the carrier at *M1* by $2\pi/N$, the carrier at *M3* lags the carrier at *M2* by the same amount, and so on up to MN whose carrier input lags on the *M1* carrier by $(N-1)\,2\pi/N$, thus giving N phases for the N paths.

The sidebands generated by these modulators will also have N phases as shown in Figure 3.

At the output of the low-pass filters *F1* through *FN* (Figure 1) the folded-difference-sideband vectors will rotate in a clockwise or anti-clockwise direction depending on whether the frequency of the voice signal is above or below the frequency of the carrier signal.

The output modulators *M1′* through *MN′* are also supplied with N phases of carrier-frequency signal f_2, each phase shift again being a multiple of $2\pi/N$.

The first translation in modulators *M1* through *MN* prepares the voice-frequency signal by folding the sidebands within the voice-frequency band and the second translation in modulators *M1′* through *MN′* puts the voice-frequency signal in its correct position in the group band.

The signals from the final modulators are now passed to a summation network. Summing the outputs of the N paths now removes either the erect or the inverted copy of the original signal, leaving only the desired output. The signal is finally passed through a broad-band filter to remove harmonics.

2.2 Mathematical Analysis

2.2.1 Notation

$f(t)$	–	function of time
$F(p)$	–	The Laplace transform of the function $f(t)$
p	–	the complex variable $j\omega$
$\omega_1 = 2\pi f_1$	–	angular frequency of input modulator carrier
$\omega_2 = 2\pi f_2$	–	angular frequency of output modulator carrier
N	–	number of paths
R_L	–	Fourier coefficient of Lth term in expansion of input modulator switching function
Q_K	–	as for R_L but for output modulator
$H(p)$	–	transfer function of low-pass filter
f_c	–	low-pass filter cut-off, $= \frac{1}{2}$ system bandwidth

2.2.2 Analysis

Consider the nth path (Figure 4). The following relations apply

$$V_2(t) = V_1(t) \cdot r(t), \tag{1}$$

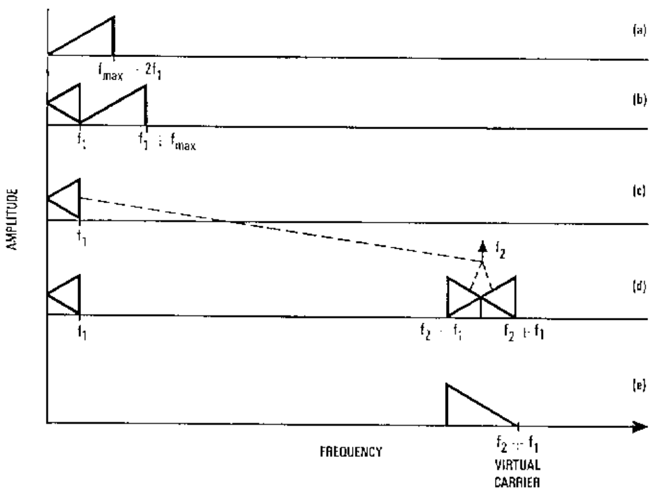

Figure 2 – Spectra at various points in the system.
(a) input
(b) output of modulator $M1$
(c) output of low-pass filter $F1$
(d) output of modulator $M1'$. The upper and lower sidebands are about carrier f_2
(e) combined outputs of N paths after band limiting

$$V_3(t) = V_2(t)\cdot h(t), \tag{2}$$

$$V_4(t) = V_3(t)\cdot q(t). \tag{3}$$

The modulating or switching functions are defined by the Fourier series

$$r(t) = \sum_{L=-\infty}^{L=+\infty} R_L\, e^{j\omega_1 Lt}, \tag{4}$$

where $R_L = \dfrac{1}{T_1}\displaystyle\int_{-T_1/2}^{+T_1/2} r(t)\, e^{-j\omega_1 Lt}\, dt,$

and $\omega_1 = 2\pi f_1 = 2\pi/T_1;$

also $q(t) = \displaystyle\sum_{K=-\infty}^{K=+\infty} Q_K\, e^{j\omega_2 Kt},$

where $Q_K = \dfrac{1}{T_2}\displaystyle\int_{-T_2/2}^{+T_2/2} q(t)\, e^{-j\omega_2 Lt}\, dt,$

and $\omega_2 = 2\pi f_2 = 2\pi/T_2$

From (1) and (4),

$$V_2(t) = V_1(t)\sum_{L=-\infty}^{L=+\infty} R_L\, e^{j\omega_1 Lt}.$$

Taking the Laplace transform

$$V_2(p) = \sum_{L=-\infty}^{L=+\infty} R_L\cdot V_1(p-Lp_1),$$

where $p_1 = j\omega_1$,

giving

$$V_3(p) = \sum_{L=-\infty}^{L=+\infty} R_L\cdot H(p-Lp_1)\cdot V_1(p-Lp_1),$$

and

$$V_4(p) = \sum_{\substack{L=-\infty\\K=-\infty}}^{\substack{K=+\infty\\L=+\infty}} R_L\cdot Q_K\cdot H(p-Lp_1)\cdot V_1(p-Lp_1-Kp_2).$$

Finally

$$V_0(p) = \sum_{n=1}^{n=N} V_4(p),$$

that is

$$V_{out} = \sum_{\substack{n=1\\L=-\infty\\K=-\infty}}^{\substack{K=+\infty\\L=+\infty\\n=N}} R_{Ln}\cdot Q_{Kn}\cdot H(p-Lp_1)\cdot V_1(p-Lp_1-Kp_2). \tag{5}$$

Consider the case where the switching function in the nth path is delayed behind the switching function in the $(n-1)$th path by a time T/xN, being in all other respects identical;

then

$$R_{Ln} = R_{L_{n-1}}\, e^{-j(2\pi L/xN)},$$
$$= R_{L1}\, e^{-j2\pi L(n-1)/xN}.$$

Similarly

$$Q_{Kn} = Q_{K1}\, e^{-j2\pi K(n-1)/xN}.$$

Now L, K, n and N are integers, and for integral values of x,

$$\sum_{n=1}^{n=N} R_{Ln}\cdot Q_{Kn} = R_{L1}\cdot Q_{K1}\sum_{n=1}^{n=N} e^{-j2\pi\left(\frac{K+L}{xN}\right)(n-1)}$$

Let $K+L = mxN$, then using the identity

$$1+a+a^2+\ldots+a^{N-1} = \frac{1-a^N}{1-a},$$

$$\sum_{n=1}^{n=N} R_{Ln}\cdot Q_{Kn} = R_{L1}\cdot Q_{L1}\,\frac{\sin(\pi mN)}{\sin(\pi m)} \tag{6}$$

ignoring the absolute phase angle.

Two cases are of particular interest; those for which $x = 1$ and $x = 2$.

Case 1 ($x = 1$)

$$K + L = mN,$$

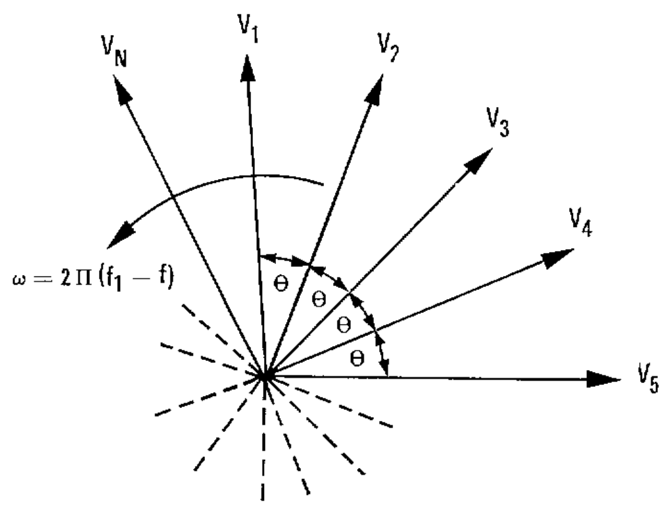

Figure 3 – Vector diagram of signals at the outputs of low-pass filters $F1$ through FN.
$\Theta = 2\pi/N$ $\qquad$ ω = angular rotation velocity

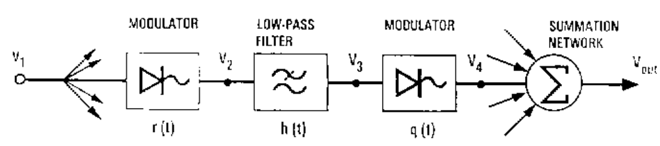

Figure 4 – Schematic representation of the nth path used in the analysis.

$$\sum_{n=1}^{n=N} R_{L_n} \cdot Q_{K_n} = N \cdot R_{L1} \cdot Q_{L1}, \text{ if } m \text{ is an integer,}$$
$$= 0 \text{ if } m \text{ is not an integer.}$$

(Note that if m is not an integer it must take a value which is the ratio of two integers).

Case 2 ($x = 2$).

This is a special case where the switching functions $r(t)$ and $q(t)$ are symmetrical functions such that they contain only odd harmonics. Only the case where K and L are both odd need be considered, so that $K + L$ = an even integer, and

$$\tfrac{1}{2}(K + L) = \tfrac{1}{2}mxN = mN$$
$$= \text{an integer.}$$

Therefore (6) will be zero unless m is also an integer, that is, when

$$\sum_{n=1}^{n=N} R_{L_n} \cdot Q_{K_n} = N \cdot R_{L1} \cdot Q_{L1}.$$

As an example, consider the multiplying function shown in Figure 5.

Here

$$R_{L1} = \frac{2}{\pi L} \sin \frac{\pi L}{2N} \sin \frac{\pi L}{2};$$

f, in addition

$$Q_{K1} = \frac{2}{K} \sin \frac{\pi K}{2N} \sin \frac{\pi K}{2},$$

the output of the system will then be

$$V_{out} = \frac{4N}{\pi^2} \sum_{\substack{L=-\infty \\ K=-\infty}}^{\substack{K=+\infty \\ L=+\infty}} \frac{1}{LK} \sin \frac{\pi K}{2N} \sin \frac{\pi L}{2N} \sin \frac{\pi K}{2} \times$$
$$\sin \frac{\pi L}{2} H(p - Lp_1) \cdot V_1(p - Lp_1 - Kp_2), \quad (7)$$

with $K + L = 2mN$, and m an integer.

Now the low-pass filters would normally be designed to transmit signals $f_1 - f$ and suppress signals $f_1 + f$. These signals correspond to the multiplier products $L = +1$ and $L = -1$. In fact only signals corresponding to $L = +1$ will be transmitted by the system since a band-limiting low-pass filter would precede it.

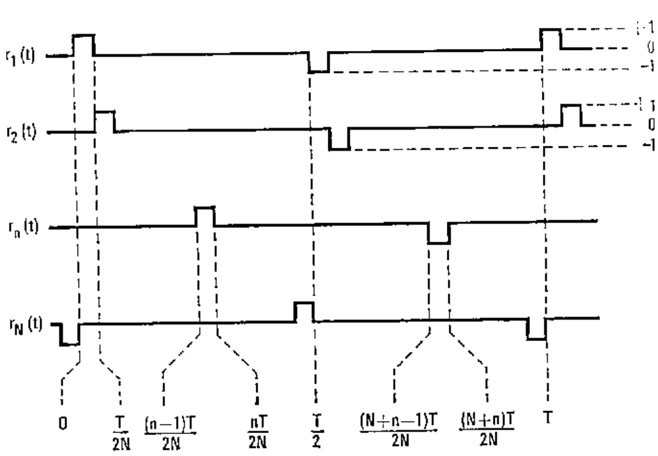

Figure 5 – Symmetrical multiplying function for N paths.

This then gives

$$K = 2mN - 1,$$

where m can take any integral value between $-\infty$ and $+\infty$.

Considering first $m = 0$; this gives $K = -1$, corresponding to an output $f_2 - f_1 + f$, which is the required single sideband. Apart from this there will be a large number of other output signals corresponding to all the other possible values of m. The signal nearest in frequency to the required output is that corresponding to $m = +1$, which makes $K = 2N - 1$ and the frequency $(2N-1)f_2 + f_1 - f$. A band-limiting low-pass filter will be required at the output to cut off all signals above $(2N-1)f_2 + f_1 - f_{max}$, where f_{max} is the maximum frequency of the input signal. This is normally less than $2f_1$. The output signal is then

$$V_{out} = \frac{4N}{\pi^2} \sin^2 \frac{\pi}{2N} \cdot H(p - p_1) \cdot V_1(p - p_1 + p_2). \quad (8)$$

The lower limit on the number of paths is set by the nearest frequency in the output $(2N-1)f_2 + f_1 - f_{max}$, which must be filtered out. If $N = 1$, this falls directly in the required output band and cannot be removed. The minimum is therefore 2 paths.

2.3 Circuit Realization

One method of realizing symmetrical switching functions is shown in Figure 6.

A1 is a phase-splitting amplifier and *A2* is a difference amplifier.

Either or both could be replaced by center-tapped transformers.

Switch *S1* alternately connects the wiper of switch *S2* to $+V_1$ and $-V_1$, dwelling on each contact for a period $T_1/2$, where $T_1 = 1/f_1$. While *S1* dwells on one contact, *S2* makes exactly one complete revolution, then *S1* changes over just as *S2* leaves the Nth contact and makes with the 1st contact. The switches are assumed to jump from contact to contact instantaneously. Switches *S3* and *S4* operate in exactly the same manner as *S2* and *S1* respectively, but with a different speed of rotation.

2.4 Supplementary Polyphase Modulation

The object of employing supplementary polyphase modulation is to remove the first unwanted product from the output of each of the input multipliers. This product is the one corresponding to $L = -1$ and is at a frequency $f + f_1$. It is the upper sideband produced by the modulation of the input signal with the fundamental component f_1 of the input multiplier function. It can be done with quadrature modulation alone or with polyphase modulation. At

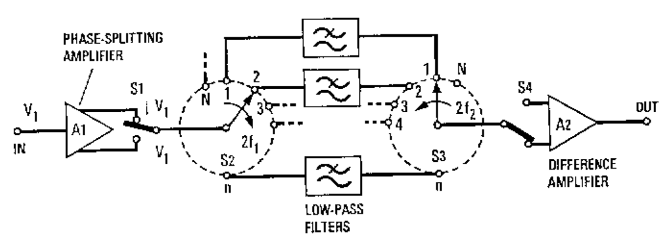

Figure 6 – Circuit realization of the switching functions in Figure 5.

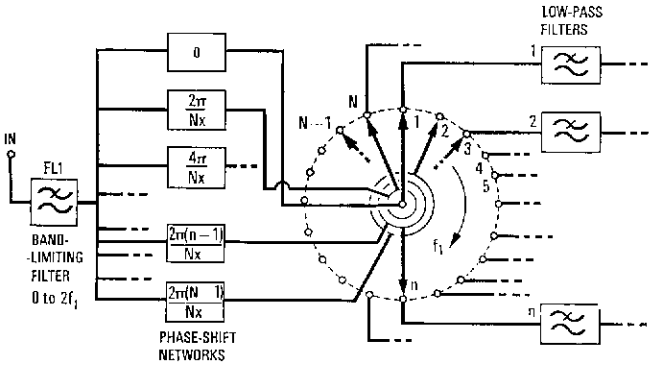

Figure 7 – Supplementary polyphase modulation.

least 30 to 40 decibels suppression may be expected by this means and the low-pass filter will be correspondingly reduced in complexity.

Figure 7 shows a circuit realization for such a modulation process. It is arranged for unbalanced modulation as the circuit for balanced modulation would be too complex to show in the general case.

2.5 General Analysis

The input voltage to the first low-pass filter (Figure 4) is given by

$$V_2 = \left[\sum_{L=-\infty}^{L=+\infty} R_{L1} \cdot V_1(p-Lp_1)\right]\left[\sum_{n=1}^{n=N} e^{-j2\pi\left(\frac{L-1}{Nx}\right)(n-1)}\right],$$

$$= N \sum_{L=-\infty}^{L=+\infty} R_{L1} \cdot V_1(p-Lp_1),$$

where, for unbalanced systems, $L-1 = m_1N$; or, for balanced systems, $L-1 = 2m_1N$, m_1 an integer.

Otherwise, for $m_1 \neq$ integer,

$V_2 = 0$.

The output voltage from the whole system is therefore

$$V_{out} = N^2 \sum_{\substack{L=-\infty \\ K=-\infty}}^{\substack{K=+\infty \\ L=+\infty}} R_{L1} \cdot Q_{K1} \cdot H(p-Lp_1) \cdot V_1(p-Lp_1-Kp_2), \quad (9)$$

with $L-1 = m_1N$, $K+L = mN$ for unbalanced systems,

$L-1 = 2m_1N$, $K+L = 2mN$ for balanced systems.

Thus, provided that N exceeds 2 for unbalanced systems and 1 for balanced systems, L can never take the value -1.

Although the system described will give sufficient suppression of the $f+f_1$ term, there will only be a limited suppression of input signals which extend beyond the range 0 to $2f_1$.

When the input frequency f is greater than $2f_1$ the difference signal $f-f_1$ will exceed f_1. This difference signal, which is the signal normally required to be passed, suffers no attenuation from the polyphase modulation and only a limited attenuation from the low-pass filters in the N paths. It is therefore necessary to insert an input band-limiting filter (*FL1* in Figure 7) such that the sum of the losses of the band-limiting filter, and that of one of the identical filters in the N paths, meets the system requirements.

2.6 Unbalancing a Symmetrical Network

If the system is disturbed from symmetry by, for example, inequalities in the low-pass filters, then certain undesirable products that would otherwise be suppressed will be present at the system output. The effects of such disturbances may be analyzed by using the theory of symmetrical components [1]. In the present context, this theory states that any unsymmetrical system of N phases may be represented by the superposition of N symmetrical N-phase systems of vectors. As an example, Figure 8 shows the resolution of an unsymmetrical 4-phase system of vectors into the sum of 4 symmetrical sets of 4-phase vectors, from which the following set of equations may be written down:

$$a_1' = a_0 - ja_1 - a_2 + ja_3$$
$$a_2' = a_0 - a_1 + a_2 - a_3$$
$$a_3' = a_0 + ja_1 - a_2 - ja_3$$
$$a_4' = a_0 + a_1 + a_2 + a_3$$

Inverting the matrix of coefficients gives:

$$a_0 = \tfrac{1}{4}\,[a_1' + a_2' + a_3' + a_4']$$
$$a_1 = \tfrac{1}{4}\,[ja_1' - a_2' - ja_3' + a_4']$$
$$a_2 = \tfrac{1}{4}\,[-a_1' + a_2' - a_3' + a_4']$$
$$a_3 = \tfrac{1}{4}\,[-ja_1' - a_2' + ja_3' + a_4'].$$

where a_1', a_2', a_3', and a_4' represent the unsymmetrical vectors and a_0, a_1, a_2, and a_3 represent 1 of the 4 vectors in each of the 4 respective symmetrical systems.

If there is an error of magnitude of 1 percent in one path, then

$$a_1' = 1, \quad a_2' = j, \quad a_3' = -1, \quad a_4' = -j0{\cdot}99;$$

therefore

$$a_0 = j0{\cdot}0025 \qquad a_1 = j0{\cdot}0025$$
$$a_2 = j0{\cdot}0025 \qquad a_3 = -j0{\cdot}9975$$

a_3 is the wanted 4-phase vector sequence and a_1 is the unwanted reverse sequence which gives rise to the unwanted superimposed inverted sideband. The discrimination is therefore

$$20 \log_{10} \left|\frac{a_3}{a_1}\right| = 52{\cdot}02 \text{ decibels},$$

in this example.

In a real measurement only the projection of a rotating vector onto a fixed datum line and the relative phase angle of 2 vectors can be detected.

3. Practical Realization

3.1 System Requirements and Design Considerations

The requirement is to translate 12 voice-frequency channels into the 60 to 108 kilohertz CCITT* basic group B and to design the system so that it is compatible with existing channeling equipment. Details depend on the requirements of individual administrations, but typical critical factors are as follows:

* International Telegraph and Telephone Consultative Committee.

— voice-frequency bandwidth 0·3 to 3·4 kilohertz.
— voice-frequency to high-frequency amplitude response $\pm 0{\cdot}25$ decibel over the band.
— Interference into adjacent channels:
Intelligible: −70 dBm0
(decibels reference 1 milliwatt at a point of zero reference level);
Non-intelligible: −60 dBm0.
— Voice-to-voice weighted noise must not exceed −74dBm0.

The optimum carrier frequency f_1 for the input modulators *M1* to *MN* is determined from the arithmetic mean of the upper (U) and lower (L) limits of the voice band, that is

$$f_1 = \frac{f_U + f_L}{2} = 1{\cdot}85 \text{ kilohertz.}$$

In practice it is convenient to adopt 1·8 kilohertz as the value of f_1 for ease of carrier generation.

The N low-pass filters must be designed to pass only the difference-frequency $f_1 - f$, and to attenuate $f_1 + f$ and all frequencies above this value by more than 60 decibels. The low-pass filter specification is therefore
— Pass band: 0 to 1·6 kilohertz $\pm 0{\cdot}25$ decibel,
Stop band: 2·1 kilohertz to $\infty > 60$ decibels.
This requires a 7th-order elliptic-function filter.

The noise requirement is a critical factor because the modulators operate with carriers at the band center, so any carrier leak will appear as an audible tone at 1·8 kilohertz. It is therefore necessary to use a special modulation circuit. The most satisfactory method to date has been to use insulated-gate field-effect transistors (*MOSFETs*) as series analog gates. Because of the high isolation between the control electrode (the gate) and the controlled path (the source-to-drain channel) it is possible to open and close the controlled path with very little leakage of the controlling signal into the controlled path.

3.2 Circuit Realization

Economic considerations alone dictate that a minimum number of 2 paths (see Section 2.2) should be used. The outline circuit is shown in Figure 9. After amplification by constant-current output amplifier *A1*, the signal is sequentially gated into 4 paths by transistors *Q1* through *Q4*. The products of modulation are filtered by balanced filters *F2* and *F3*.

Because the filters use balanced coils no attenuation is offered to input signals which are not in 4 phases. It is therefore necessary to include the coaxial chokes *X1* and *X2* to suppress residual unwanted output components. The 4-phase difference signal $f - f_1$ that appears at the filter outputs is sampled sequentially by *Q5* through *Q8* and the resultant signal, containing only the single sideband plus harmonics, is passed through the buffer amplifier *A2*.

In this way 12 channels are generated, and after combining are passed through the common group filter *F4* which removes harmonic sidebands.

3.3 Signaling

The insertion of low- or high-level outband signaling at 3825 hertz may be achieved simply.

A master oscillator provides a signal at 2025 hertz which is keyed by the signaling information in the usual way. This signal is then filtered and split into 2 identical signals in phase quadrature which are injected by a separate winding into the last coils of filters *F2* and *F3*. In this way the equivalent of a keyed 3825-hertz outband signal is inserted. Insertion at voice frequency would require increased voice-frequency low-pass filtering as well as a filter with a pass band at 3825 hertz. In a conventional system the method used would be equivalent to inserting the signaling at high frequency after the channel band-pass filter.

3.4 Demodulator

The operation of the demodulator voice and signaling paths is similar to that of the modulator, but in reverse. Because each channel handles the complete 12-channel signal power, care has to be exercised in the circuit design to prevent intermodulation distortion.

3.5 Modulator Performance —
(104- to 108-kilohertz channel).

3.5.1 Gain Stability

— Temperature: $< 0{\cdot}2$ decibel for any 20 degrees Celsius rise in the range 0 to 40 degrees Celsius (over entire frequency range).

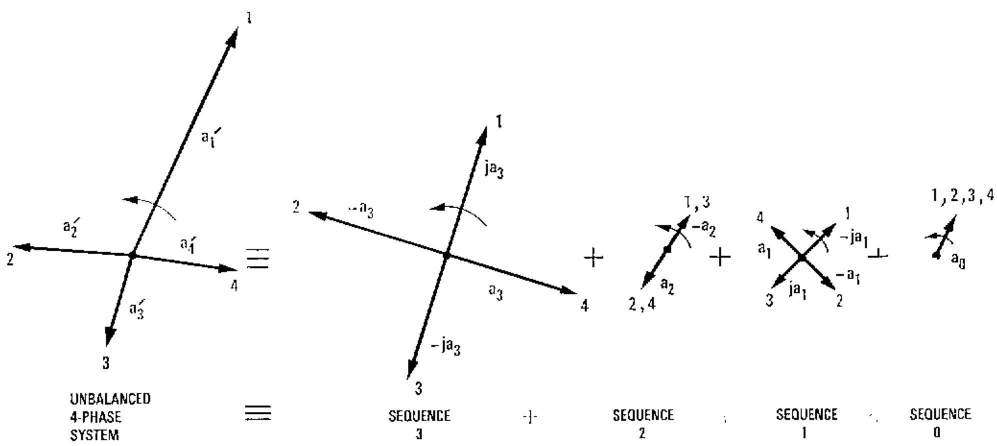

Figure 8 – 4-phase symmetrical components. The numbers at the end of each vector refer to the paths in which the vectors are observed.

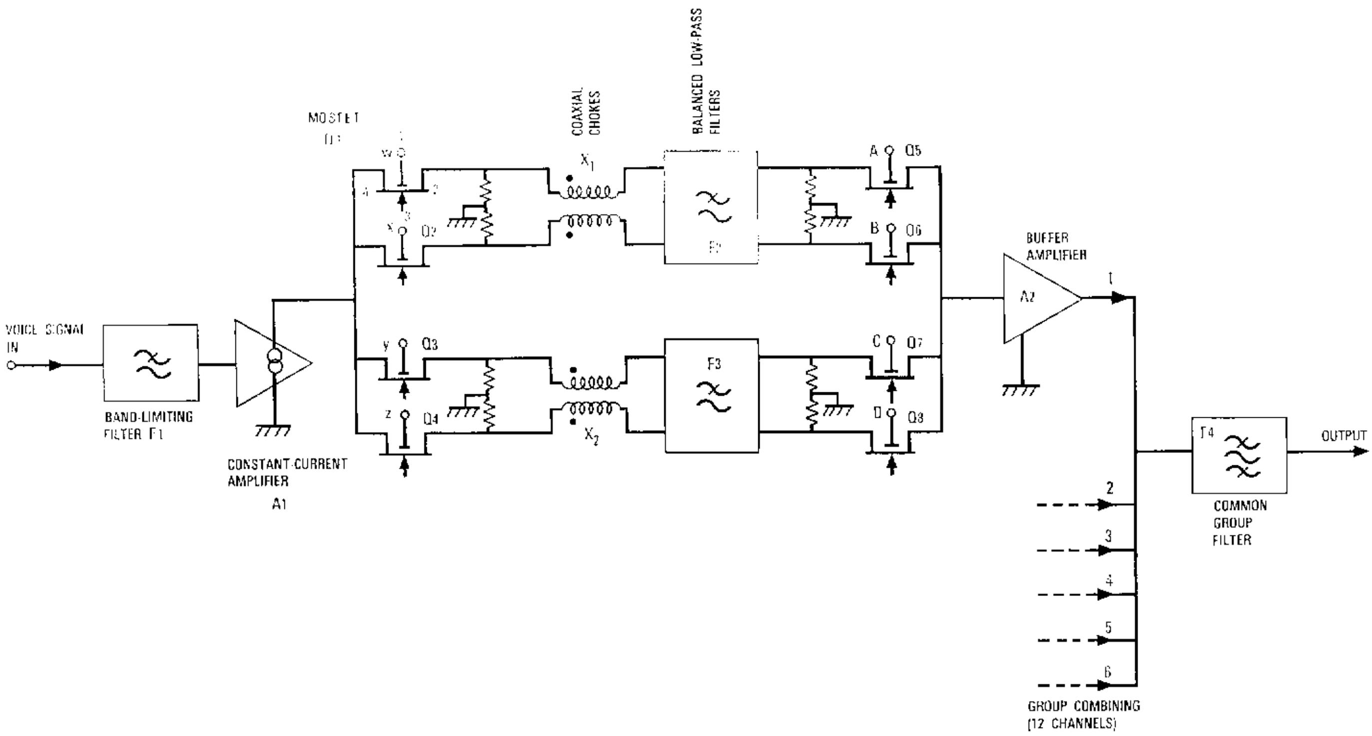

Figure 9 – Schematic showing transmit side of 1 channel and group combining. Q_1 through Q_8 are metal-oxide-silicon field-effect transistors (*MOSFETs*). Looking at Q_1 the electrodes are arranged as follows

1 – gate *W* *2* – source *3* – unconnected substrate *4* – drain

The gate inputs to the other *MOSFETs* are *X, Y, Z, A, B,* C, and *D*. In all cases the substrate *3* is unconnected.

— Supply voltage: $< \pm 0{\cdot}05$ decibel for $\pm 1{\cdot}5$ percent supply change.

— Carrier supply: Reasonably independent of carrier supply level.

3.5.2 *Frequency Response*

Voice-to-high or high-to-voice bands

250 hertz to 3·15 kilohertz	$> -0{\cdot}5$ decibel $< +0{\cdot}2$ decibel	referred to an 800-hertz test tone
3·15 to 3·40 kilohertz	> -1 decibel	

3.5.3 *Linear Region*

With the transmit channel output looped back to the receive channel input, the voice-to-voice response does not deviate from linearity by more than $\pm 0{\cdot}3$ decibel for an increase in signal level from -12 dBm0 to ± 4 dBm0 (at 800 hertz).

3.5.4 *Virtual Carrier Leak* ($f_2 \pm f_1$)

$f_2 \pm f_1 < -70$ dBm0 (at 104·4 and 108·0 kilohertz).

3.5.5 f_2 *Carrier Leak*

Level of f_2 (106·2 kilohertz) is -80 dBm0.

3.5.6 *Noise*

The noise level is primarily limited by the 1·8-kilohertz tone because of carrier leak and is therefore -80 dBm0 unweighted.

3.5.7 *Sideband Interference*

— Linear products: These are limited by the suppression available in the low-pass filter. Unintelligible crosstalk is -65 dBm0. The worst intelligible crosstalk is -72 dBm0.

— Non-linear products: Because of distortion in the output modulators and buffer amplifier, certain unintelligible products appear in adjacent channels. The largest is -62 dBm0.

4. Conclusions

It has been demonstrated practically that unwanted products of modulation and carrier leak can be controlled in this system to meet the general channeling requirements. Measurements have been made on a model with *LC* filters to demonstrate feasibility and a suitable design of active filter has been developed. It is concluded that the objective of a completely integrated channel translator can be attained.

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Patents

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